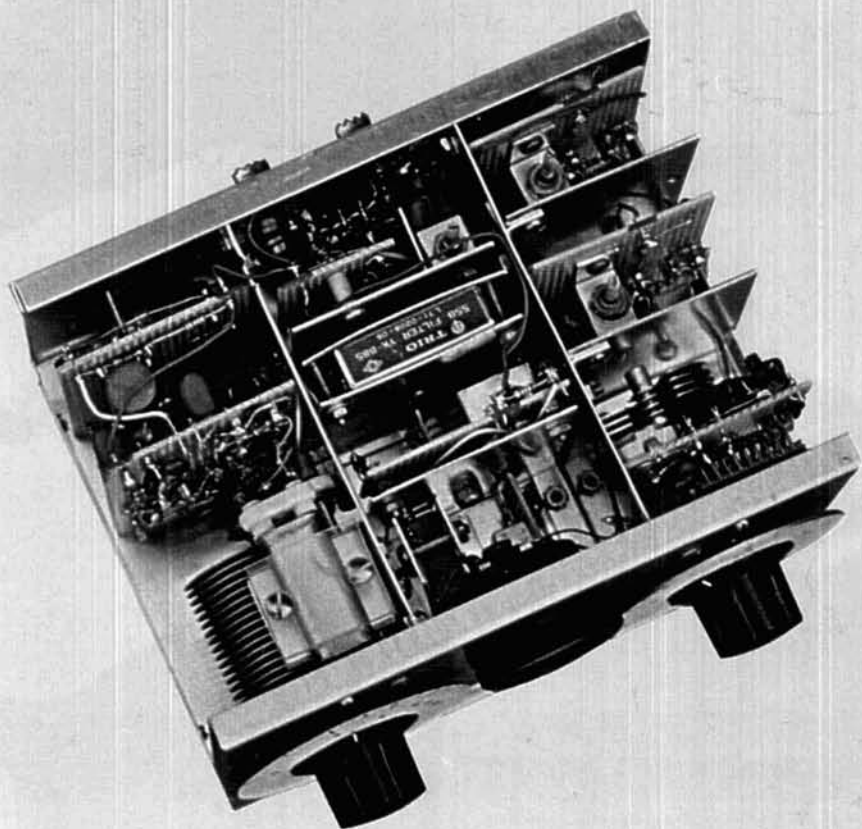


ham radio magazine

- HF receiver performance
- better-sounding SSB
- meteor-scatter communications
- designing elliptic lowpass filters
- EMI/RFI shielding: new techniques (part 2)

*THE QUIET FREQUENCY SYNTHESIZER:
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150

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TS-930S

"DX-traordinary" ... superior dynamic range, auto. antenna tuner, QSK, dual NB, 2 VFO's, general coverage receiver.

A superlative, high-performance, all solid-state HF transceiver, that covers all Amateur HF bands, and incorporates a 150 kHz to 30 MHz general coverage receiver having an excellent dynamic range.

TS-930S FEATURES:

- 160-10 Meters, with 150 kHz-30 MHz general coverage receiver. Covers all Amateur frequencies, plus WARC, on SSB, CW, FSK, and AM. UP conversion digital PLL circuit.
- Excellent receiver dynamic range. Typical two-tone dynamic range, 100 dB (20 meters, 50-kHz spacing, 500 Hz CW bandwidth).
- All solid-state 28 volt operated final amplifier. Lowest IM distortion. Power input 250 W on

- SSB/CW/FSK, 80 W on AM, SWR/ Power meter.
- Available with AT-930 automatic antenna tuner built-in, or as an option. Covers 80-10 meters, including WARC bands.
- CW full break-in. CMOS logic IC, plus reed relay. Switchable to semi break-in.
- Dual digital VFO's, 10-Hz steps, includes band information.
- Eight memory channels. Stores frequency and band data. Internal battery memory back-up, est. 1 yr. life. (Battery not Kenwood supplied.)
- Dual mode noise blanker, NB-1, with threshold control, for "pulse" noise. NB-2 for "woodpecker."

- SSB IF slope tuning, allows independent adjustment of the low and/or high frequency slopes of the IF passband.
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- Tuneable, peak-type audio filter for CW.
- AC power supply built-in.
- Fluorescent tube digital display (100 Hz resolution, modifiable to 10 Hz) with digitalized sub-scale, in 20-kHz steps.
- RF speech processor.
- One year limited warranty.

- SSB monitor circuit.

Optional Accessories:

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"Digital DX-terity" ... General coverage, Superior dynamic range, 2 VFO's, 8 memories, Scan, Notch, COMPACT!

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- Narrow-wide filter selection on SSB and CW (filter optional).
- Speech processor, built-in.
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- Fluorescent tube digital display indicates frequency to 100 Hz (10 Hz modifiable).
- All-mode squelch circuit, built-in.
- Built-in noise blanker.
- RF attenuator (20 dB).
- VOX circuit, plus semi break-in with side-tone.

Optional accessories:

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- SP-430 external speaker.
- MB-430 mobile mounting bracket.
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- YK-88A (6 kHz) AM filter.
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cover: W3MT

February 1984  5

REFLECTIONS

I enjoy operating because it's relaxing (sometimes), informative ("Roger, you're 5 and 9 — NEXT!"), and a good way to meet interesting people (that's true). It gives me an opportunity to try out new antenna designs, test my understanding of propagation, and, in general, stay more or less informed about what's happening. However, in the process I've noticed a trend worth examining — particularly by those who are part of it. I call it:

INSTANT PUDDING

Recently I've found myself reduced to mumbling into my beard about such-and-such practice by this-or-that individual. It appears that technically, at least, Amateur Radio has come of age, with instant power-on, QSY, tune-up, and antenna matching. We have the latest fully-synthesized, all-mode transceivers, full break-in amplifiers, and motor driven crank-up towers. With even half an effort we can work DXCC in a couple of weeks (or even on a weekend, during a contest). But in the process, something's been left behind. Call it courtesy, experience, operating skill, or just plain consideration for our fellow hams. Let me give you a few examples of what I'm talking about.

WD22XYZ calls CQ, is answered by a DX station, completes the contact, and stands by for any other DX calls on his frequency. Without a moment's hesitation, KD11PQC jumps in with kilowatts a-glowing and calls the DX station. When I was a novice to the hobby I learned fast that this is simply not done; unless WD22XYZ relinquishes the frequency, it's still his.

Another pet peeve of mine, heard more times than I care to recall, occurs when a DX station calls a directional CQ: "CQ, CQ, CQ, listening for W1's only!" The ensuing cacophony of calls includes all areas, states, and cities (with only slight exaggeration on my part). Of course, the DX station can control the situation if he continues to request the particular call area and refuses to give in by answering that loud W12 — *maybe*.

Signal quality reporting is another troublesome area. God help the operator who informs another that his signal leaves a little bit to be desired; for example, 20 kHz wide on glorious SSB. I made the mistake of telling another ham the other day that he was transmitting on two different frequencies simultaneously. His immediate reaction was, "That's impossible! It's probably your receiver that's overloading." I hastened to add that *both* of his signals were quite "clean"; it's just that he had two VFO's going at the same time. In all fairness to him, it turned out to have been the other operator in the QSO who was at fault. Fellas — I think most criticism is meant to be constructive. (Anyhow, whatever happened to pride in one's signal/station?)

How many times have you heard, "Roger, roger, roger, you're 5-9 plus. But will you please repeat your call and name and my report?" Please note that Q5 means "perfect copy" and S9, by IARU convention, is 50 microvolts RMS (in a 50 ohm system). By the way, it might be very useful to everyone if you knew what your Brand X receiver S9 was equivalent to. It might make the signal reports a little more meaningful when testing antennas or evaluating the performance of one's station over another's.

Full-power tune-up into the antenna seems to be occurring more and more often these days, usually when you're trying to pull that weak one out. If you *must* tune up on frequency (I won't mention dummy loads), listen first, wait, listen again to make sure that it's a clear frequency and *then* give a short request to see whether the frequency is in use. If it isn't, go ahead. Propagation being what it is, it's quite possible to believe that a frequency's clear when it's really quite occupied. If you have to apply full power to your antenna, load up when the band is truly dead: high noon on 160 meters, 4 A.M. on 10 meters. (I know — 160 and 10 meter operators will insist the bands are *never* dead.)

Communicasting is the term I use to identify operation by an individual who wants others to think he's communicating, even though he's really only broadcasting his highly opinionated, usually insulting, narrow-minded point of view. This practice has reared its ugly head more and more often on the lower bands recently — all in the name of exercising the right to free speech. With all due respect to rights, isn't use of the frequency spectrum — a precious, limited resource — a *privilege*? (If you have something to say, and you want other Amateurs to hear it, write a letter to *ham radio*. If space permits, we'll print it.)

To paraphrase a recent television ad describing the reasons for a stock brokerage's success, courteous operation is just "good old-fashioned work." In our case, if we were to stop expecting instant pudding — i.e., immediate gratification — and instead were more willing to hang in there, learn by example, develop expertise and the kind of operating skills that can only be acquired over a period of time, I, for one, would be a lot happier.

Rich Rosen, K2RR
editor-in-chief

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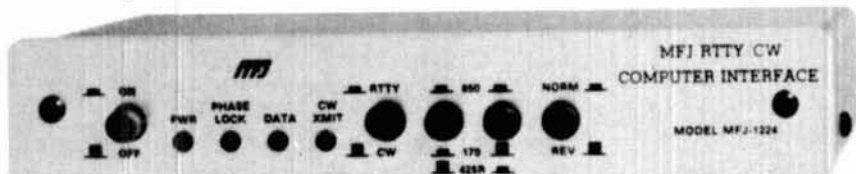
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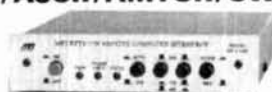
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THE "NO-CODE" LICENSE PROPOSAL WAS SOUNDLY DEFEATED, with not one commissioner voting for it at the December 14 FCC meeting. Presenting the item was Private Radio Bureau Chief Bob Foosaner, who pointed out that while a principal argument for No-Code was the supposed lack of growth in Amateur Radio, the service had grown from 300,000 licensees in 1975 to over 400,000. Furthermore, he continued, the nearly 5,000 comments filed ran 20:1 against No-Code, with even the handicapped it was supposed to benefit opposing it.

Commissioner Mimi Dawson, Studying For An Amateur License Herself, then asked what alternative entry to the radio spectrum might be available for "computer hobbyist" radio communications since Amateur opposition to No-Code did not make Amateur Radio attractive to that group. Foosaner responded, "The hobby class (license) may be something that replaces the CB type of thing...we'll find a spectrum available for this type of service."

Commissioner Comments During The Discussion Were All Complimentary to Amateur Radio, with Chairman Fowler in particular citing the service's many contributions. However, some observers feel the failure of No-Code and the resulting lack of growth in Amateur VHF or UHF band usage will greatly increase outside pressure on those bands. Not only is there the possibility of a non-Amateur "hobby class" competitor for spectrum, but various user groups are actively promoting 216-225 MHz as a land mobile band, plus use of 420-430 MHz.

W5LFL'S STS-9 2-METER OPERATION WAS AN UNQUALIFIED SUCCESS, from every point of view. He operated enough orbits to give the U.S. and the rest of the world at least several chances to hear and—hopefully—"work" him. His "CQ North America" was easily copyable. a thousand miles from Columbia's groundtrack, depending on spacecraft orientation.

Media Coverage Of The Event Was Outstanding, emphasizing the idea that now "the guy next door" could actually talk to an astronaut in space. Owen's chat with King Hussein, JY1, was highlighted, of course, but many average Amateurs across the country found themselves hosting TV news crews and chatting with talk show hosts or newspaper reporters.

Though Owen's Operating Method Disappointed Many who'd hoped to hear their calls come back from space, it made real PR sense. His acknowledging a few calls, then describing what he saw and what was going on in Columbia was far more interesting to non-Amateur listeners and media than long lists of calls. He also reported it almost impossible to copy many calls because of his single-earpiece headset. Though it appears most of those Owen was acknowledging were well-equipped VHF buffs with big signals, Owen's first review of his logging tape after Columbia's return did turn up some mobiles! Discipline was generally fair to good in most areas, though with some confusion and harassment.

Strong, Positive Effects On Amateur Radio are already being felt. Reinforcing the excellent exposure Amateur Radio got during the Grenada invasion, interest in becoming an Amateur seems to be at an all-time high. Inquiries about Amateur training courses are way up, and school-age Amateurs report many classmates asking about their hobby. NASA was also very pleased, with future operation by other licensed astronauts almost certain!

About 300 Callsigns Were Pulled Off The Logging Tape in Owen's initial runthrough. He feels quite a few more can come from a more painstaking review. Of the 300, about 75% were U.S., with no Japanese, Russians or Africans noted. Special QSLs for those who made it on the tape will be available shortly. Send QSLs and SWL reports to ARRL, with an SASE.

Congratulations to NASA, The ARRL, W5LFL, K6DUE, and many other contributors toward making Owen's space operation possible. A special salute to Vic Clark, W4KFC, who passed away one week too soon to see a dream on which he'd worked so hard finally realized. A most ambitious project, meticulously executed, for the benefit of all—but particularly Amateur Radio.

PCB-FILLED DUMMY LOADS DO POSE A THREAT TO AMATEURS, though perhaps not as severe as once feared. A just-completed study by the Centers for Disease Control in Atlanta found less than 2% of the loads tested contained PCB coolant, though a few others did show traces of the dangerous compound. However, tests of the Amateurs who'd been exposed to the PCB-cooled loads showed no abnormal levels in their blood.

Since Any Exposure To PCBs Is Considered Dangerous, Amateurs who have dummy loads, capacitors, or transformers made before PCB's dangers became known should contact the nearest office of the EPA for advice on safe disposal.

THE VOLUNTEER EXAM IS MOVING AHEAD, with appointment of the first regional Volunteer Exam Coordinator likely very soon. Leading the race with a very well thought out proposal is the Anchorage Radio Club, which is anxious to take on the responsibility for the entire state of Alaska. There are also groups or organizations in almost every other part of the country who've indicated strong interest in taking on the VEC job in their areas. In the meantime, the ARRL is still talking about becoming the national VEC.

WARC FREQUENCY ALLOCATIONS WERE ADOPTED BY THE FCC in December, with several items of concern for Amateurs. 1.9-2.0 MHz was allocated to Amateur Radio on a "restricted secondary" basis, with radiolocation primary. AM broadcast will eventually be moving into 1605-1705 kHz, with displaced radiolocation users then going to 160's top end. 220-225 MHz retains Amateur, plus fixed and mobile pending the results of an FCC NTIA working group study. The 902-928 MHz band seems to have a variety of potential users, but the FCC did turn down industry pleas that it also include a secondary land-mobile allocation.

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FEATURES

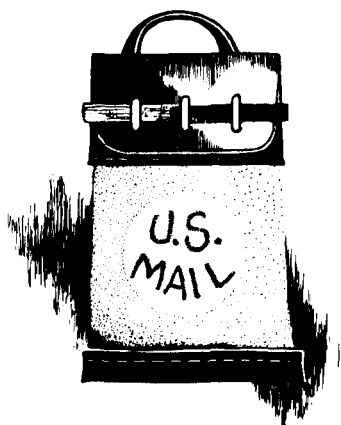
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comments

photovoltaics

Dear HR:

WD8AHO's article, "Photovoltaic Cells: a Progress Report" (December, 1983, page 52), brought to mind painful recollections of the early 1960s when I bought 300 0.4×0.8 silicon cells surplus for fifty bucks. I say "painful" because I soldered the little buggers in a series-parallel arrangement. I thought I would go blind and also learned that the little pieces of silicon would fracture easily. When most of the 300 were finally soldered, I had the pleasure of getting all of 2 + watts output on a sunny day. Still, it was great fun after the drudge work of putting it together was done and it did make a nice battery charger. After several years of operation, the unit busted up on my move from San Francisco to Oakland. Too bad — there are more sunshine/hours in Oakland.

The 'breeder concept' photo in the article (fig. 1) made me whistle. There must be at least 3000 square feet of panels. That would mean about a million of my 0.4×0.8 chips (ouch), and at 1960's prices, a cost (even at surplus) of over \$150,000, all for about 6kW output!

Since I've been out of touch with that scene for over twelve years, it's apparent there has been a great deal of progress.

Your article really rekindled my interest.

Nubar Tashjian, K6KVX
Oakland, California

Dear HR:

I don't understand WD8AHO's claim that solar cells at \$4.00 per watt will pay back in less than one year. (See "Photovoltaic Cells: A Progress Report," December, 1983, page 52.)

Energy from our local electric company costs $\frac{1W}{1000} \times \$0.052 = \0.52×10^{-4} each hour at the 1 watt rate. So, for the \$4.00 cost of a 1-watt solar cell, I could buy 76,923 hours of electric energy at the 1-watt rate. Averaging six hours per day, as one might use a solar cell, I could buy that power from the local electric company for 12,820.5 days or about 35.1 years.

The added expense for storage batteries (and inverters?) makes the break-even time even longer. I don't know the life expectancy of solar cells, but I would be surprised to break even within the span of human lifetime.

Martin Sample
Tuolumne, California

As Mr. Sample suggests, the question of when photovoltaics will become cost competitive with local power companies is, to some, a good reason not to get involved in photovoltaics.

Photovoltaics is not for everyone, and certainly not for anyone who measures payback solely in economic terms. But a modest photovoltaics system can meet the needs of the Amateur who wants an economical, uninterrupted supply of power for his or her ham shack, for remote operations, and in emergencies or natural disasters.

Every day we read of billion-dollar cost overruns, unexpected delays, and unacceptable workmanship in the development of new power plants. These problems suggest that in the years to come, electricity is likely to cost more, not less. It's reasonable to expect that by the time photovoltaic power reaches the 100 megawatt level and the cost has dropped to below \$4 per watt, the payback period of PV's — partly because of power company rate increases — will be reduced to less than one year and be competitive with utility rates.

With a life expectancy of 30 years or more, photovoltaics today offer a readily

available source of dependable power at a reasonable price.

Paul J. DeNapoli, WD8AHO
Livonia, Michigan

As far as Amateurs today are concerned, perhaps the following letter says it best . . .

Dear HR:

I fail to see how solar energy can have a payback of less than three and one-half years. At \$10 per peak watt, and four hours per day peak sun time, I calculate only 1460 watt hours per year. At 7 cents per kWh, this equates to a payback of 10 cents per year — which would take 100 years!

WD8AHO is clearly talking about energy payback only (the energy required to build a cell), not the cost of manufacture. I am truly excited about the prospects of solar energy and finally put cells on my sailboat to keep the batteries trickle-charged. You can't beat PVs when you can't get power any other way, but let's be honest about the cost to the consumer.

R.F. Bruninga, WB4APT
Bowie, Maryland

auto-dialer chip

The MD-22 chip necessary for construction of K2MWU's "State-of-the-Art Auto-Dialer" (December, 1983, page 21) will again be available in mid-to late February.

Readers who have had difficulty in obtaining this part should contact CES, Inc., P.O. Box 2930, Winter Park, Florida 32790. (305) 645-0474.

short circuit

RTTY and the Atari™

In the article, "RTTY and the Atari™ Computer," on page 38 in the July, 1983, issue of *ham radio*, the lower left pinout of the XR2211 chip should be labeled 12. It was omitted from fig. 3 though discussed in the text on the following page under the subhead "terminal unit construction adjustments."

AEA Brings You The AMTOR Breakthrough

We are pleased to announce three new AMTOR products. Our new software package that will allow you to operate AMTOR with your CP-1 is called AMTORTEXT™. A complete hardware terminal unit and AMTORTEXT software plug-in cartridge for the Commodore 64 computer is called the MICROAMTOR PATCH™. We also have new applications software packages for the AMT-1 and Commodore 64 or VIC-20 computers.

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The AMTOR software TIMING ROUTINES have been written by Peter Martinez, G3PLX (father of AMTOR) which means you can be sure of having NO SYNCHRONIZING problems with other AMTOR stations adhering to the established international AMTOR standard. PROPER SYNCHRONIZATION is an ABSOLUTE must for AMTOR!



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frequency synthesis by VXO harmonic selection

Quieter than a PLL,
it generates all frequencies
between 5 and 6 MHz

This article describes a modification of an old method of frequency synthesis.¹ Its chief advantage over the earlier scheme is that it can synthesize all possible frequencies over the output range, rather than being limited to a set of discrete frequencies. Its stability, determined by the stability of a single VXO, is comparable to that of other synthesizers.

block diagram

The basis for the system is a variable crystal oscillator — a VXO consisting of a 100-kHz frequency standard crystal which is continuously “pulled” from 100 to 100.2 kHz (see fig. 1). The VXO output frequency is divided by ten and fed to a pulse generator. Here a short pulse is formed with a PRF continuously variable over the range 10 to 10.02 kHz, directly controlled by the VXO. A bandpass filter passes the 500th through 600th harmonics of the pulse, producing a “comb” with about 10 kHz spacing which moves continuously upward about 10 kHz as the VXO frequency is increased. If the 500th harmonic is selected, its frequency will move continuously from 5000 to 5010 kHz. Then if the 501st harmonic is selected, its frequency will move continuously from 5010 to 5020.02 kHz and so on. In this way we may obtain all possible frequencies in the range from 5 to 6 MHz.

In order to select one harmonic and reject all others, it is necessary to employ a filter having a bandwidth considerably less than 10 kHz. Consideration of spurious frequency problems leads to the use of a SSB crystal filter in the 9 MHz region. I selected an 8.83 MHz filter because I intend to use a 9 MHz beating oscillator to heterodyne the synthesizer into the ham bands, and I do not wish to invite trouble. The VFO and first mixer form a superhet with an 8.83 MHz IF sharp enough to pass just one “tooth” of the “comb.” The IF beating against the same VFO in the second mixer restores the selected “tooth” to its original frequency, *exactly*. Drift in the VFO has no effect on the output frequency. As the VFO is tuned continuously in one direction, succeeding “teeth” are selected, interspersed with regions of zero output. The VFO is calibrated in terms of output frequency so that the appropriate “tooth” is selected. The output frequency is then adjusted to the desired value by adjusting the VXO frequency. If this adjustment causes the IF to reject the “tooth,” slight retuning of the VFO will bring it back into the filter pass-band. The selection indicator is a meter which indicates whether or not a “tooth” is passing through the filter. It reads either zero or about 2/3 scale, corresponding to the absence or presence of an output signal, respectively.

theory of operation

Assume a VXO is continuously “pulled” from f_0 to a higher frequency, f_1 . We may multiply the frequency by any integer, M , and divide the frequency by any integer, N . The output frequency will vary from

$$f_L = \frac{M}{N} f_0 \quad (1)$$

to

$$f_H = \frac{M}{N} f_1 \quad (2)$$

By Frank Noble, W3MT, 10004 Belhaven Road,
Bethesda, Maryland 20817

For continuous coverage, the pulling range of **fig. 2(A)** must be equal to the separation between adjacent harmonics of $\frac{f_0}{N}$, **fig. 2(B)**, or

$$\frac{M}{N} (f_1 - f_0) = \frac{f_0}{N} \quad (3)$$

so that
$$M = \frac{f_0}{f_1 - f_0} \quad (4)$$

Substituting **eq. 4** in **eq. 1**:

$$N = M \frac{f_0}{f_L} = \frac{f_0^2}{f_L (f_1 - f_0)} \quad (5)$$

Comparing **fig. 2(C)** with **fig. 2(A)**, note that

$$\frac{M + 1}{N} (f_1 - f_0) \text{ is larger than } \frac{M}{N} (f_1 - f_0) \quad (6)$$

so M in **eq. 4** is the *minimum* value which will provide continuous coverage. Larger values will produce progressively increasing overlap.

practical considerations

If a standard SSB crystal filter is used, the harmonic spacing must be of the order of 10 kHz to allow adequate rejection of neighboring harmonics. Hence

$$\frac{f_0}{N} = 10^4 \quad (7)$$

and
$$N = f_0 \times 10^{-4} \quad (8)$$

Substituting **eq. 8** in **eq. 5**:

$$M = f_L \times 10^{-4} \quad (9)$$

Now the desired value of f_L is 5×10^6 , so that

$$M = 500 \quad (10)$$

AT-Cut crystals will not achieve values of M as small as 500, using capacitance tuning. This amount of pull may be obtained with inductive tuning, but extreme care must be taken, especially with the coil, to retain stability. The frequency, f_0 , must exceed 10 MHz so that reasonable values of inductance may be used. This leads to the requirement of large frequency division, N , an additional complication. An incidental advantage is that the tuning linearity is better than for capacitance loading.

If we use the 100 kHz 5 degree X Cut commonly employed for frequency standards, a value of $M = 500$ may be obtained with capacitance tuning. The temperature stability will be poorer than for AT crystals, but a crystal oven will solve this problem.

In any case, the percentage drift of the output frequency will be identical to the percentage drift of the crystal.

circuit description

Fig. 3 shows a detailed circuit diagram for one form of the synthesizer.

The first stage, the VXO, uses an LF353N for U1A in a power design² and substitutes a large variable capacitor having "midline" plates to improve the tuning linearity. (SLF or even more radical plate shapes would be preferable if available.) A fast-rising square wave is formed by U1B, suitable for driving the divide by ten IC, U2. The very fast rising square wave output of U2 is differentiated by a short time-constant RC circuit. The negative-going pulse is clipped by the

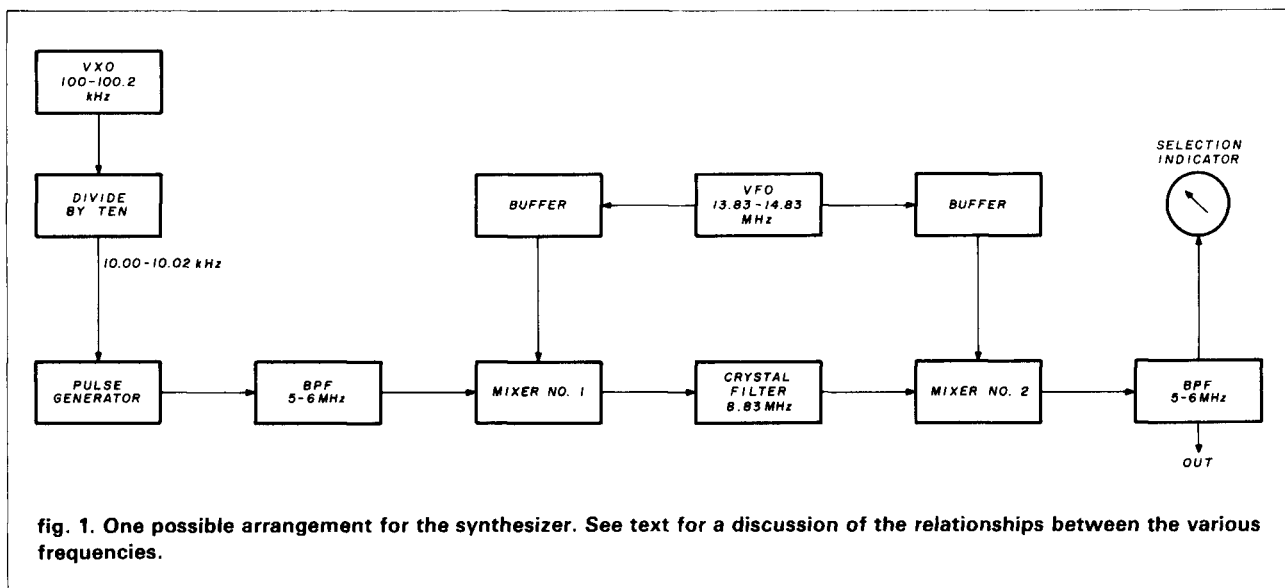


fig. 1. One possible arrangement for the synthesizer. See text for a discussion of the relationships between the various frequencies.

diode, so that the input to Q1 is a short positive pulse with a PRF continuously variable from 10 to 10.02 kHz. The source of Q1 is biased at +6 volts so that only the narrow part of the pulse near its peak is amplified. The tuned circuit in the output of Q1 is resistively loaded to pass 5-6 MHz, selecting the desired "comb." Q2 raises the comb level. Its output is also broadbanded by resistive loading. A link on the Q2 output coil drives one input of diode ring mixer No. 1 through an attenuator which provides a 50-ohm source for all frequencies. The second input to the mixer is supplied by the VFO buffer, Q9, having a source impedance of about 50 ohms at the VFO frequency. (This is not ideal. It would have been better to increase the output of Q9 and use an attenuator to provide 50 ohms at all frequencies.)

The output of the first mixer is terminated in 50 ohms for all frequencies. *This is the most critical termination.* It drives FL-1 through a series resistor, forcing a termination of less than 600 ohms, which is roughly what the filter requires. The output termination is about 400 ohms. (We observe that passband flatness, vital to SSB, is not important in this application; consequently, termination requirements have been eased.) Since the filter is very sharp, the output circuit of IF amplifier U3 has high Q and large L/C to achieve high gain. Q3 is an RC amplifier for driving FL-2 with a source impedance of somewhat less than 680 ohms. This filter is also terminated in somewhat less than 680 ohms. It drives emitter-follower Q4.

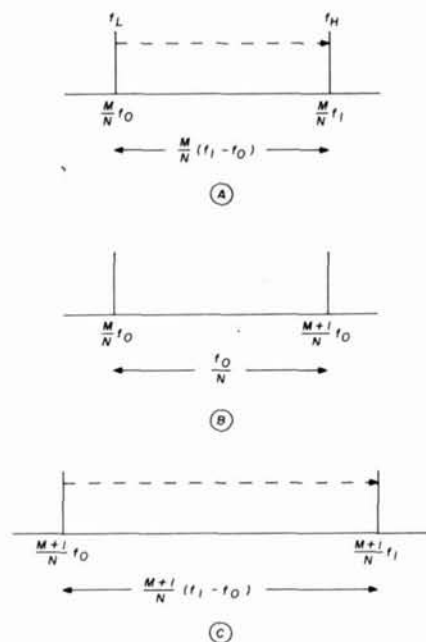
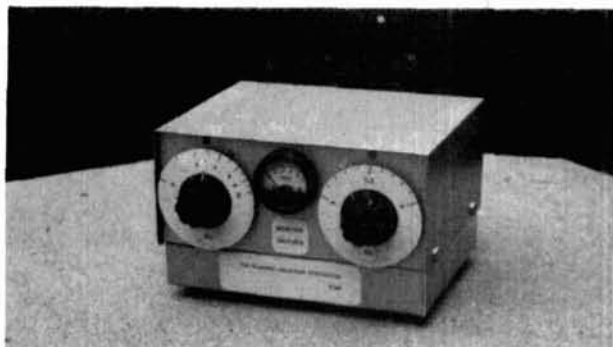


fig. 2. Diagram for determining the relationships between frequencies, multipliers, and divisors.



Left, VXO direct drive dial calibrated in kHz. Center, selection indicator meter. Right, VFO reduction drive dial calibrated in 100 kHz steps.

The output impedance of Q4 approximates 50 ohms at all frequencies to match one input of the second mixer. A separate VFO buffer, Q8, drives the other input. The critical output of the second mixer sees 50 ohms at all frequencies. U4 is an amplifier, broadbanded by the 50 ohm half-wave filter at its output. The input to amplifier U5 is fairly clean. The signal is made larger by U5 and further laundered by the bandpass filter at its output. Q5 is an emitter follower to reduce the output impedance to about 50 ohms at the "HI" output. An L network followed by a quarter-wave filter reduces the output voltage and impedance and further improves the waveform at the "LO" output. Q6 is another emitter follower used to prevent distortion generated by the diode rectifier from reaching the RF outputs. The rectified and filtered output of Q6 is used to drive the selection indicator, a miniature 1 milliamper meter. The VFO, Q7, is a Hartley oscillator using a good coil and source degeneration to obtain good stability and waveform. It must not drift over ± 1 kHz per hour — otherwise it could drift the desired "tooth" out of the filter slot. The VFO tuning capacitor has semi-circular plates for good tuning linearity. It is driven by a reduction dial to make the tuning less critical.

mechanical details

Except for the portion between the U4 link and the outputs, the circuit is contained in a metal enclosure (standard chassis No. 1) measuring 3 1/4" H \times 7 1/2" W \times 5 7/8" D, (9 cm \times 19 cm \times 15 cm) Radio Shack No. 270-229 (see photo 2). A second standard chassis (No. 2) contains the portion after the U4 link, a 12-volt regulated power supply (150 mA), and a converter employing a 12.5 MHz crystal oscillator to beat the synthesizer output up into the 40-meter band. No attempt was made to miniaturize chassis No. 2 because it rests on a shelf under the operating table, out of the way. The coils in chassis No. 2 are surplus 5/8-inch slug-tuned bakelite units mounted in shield

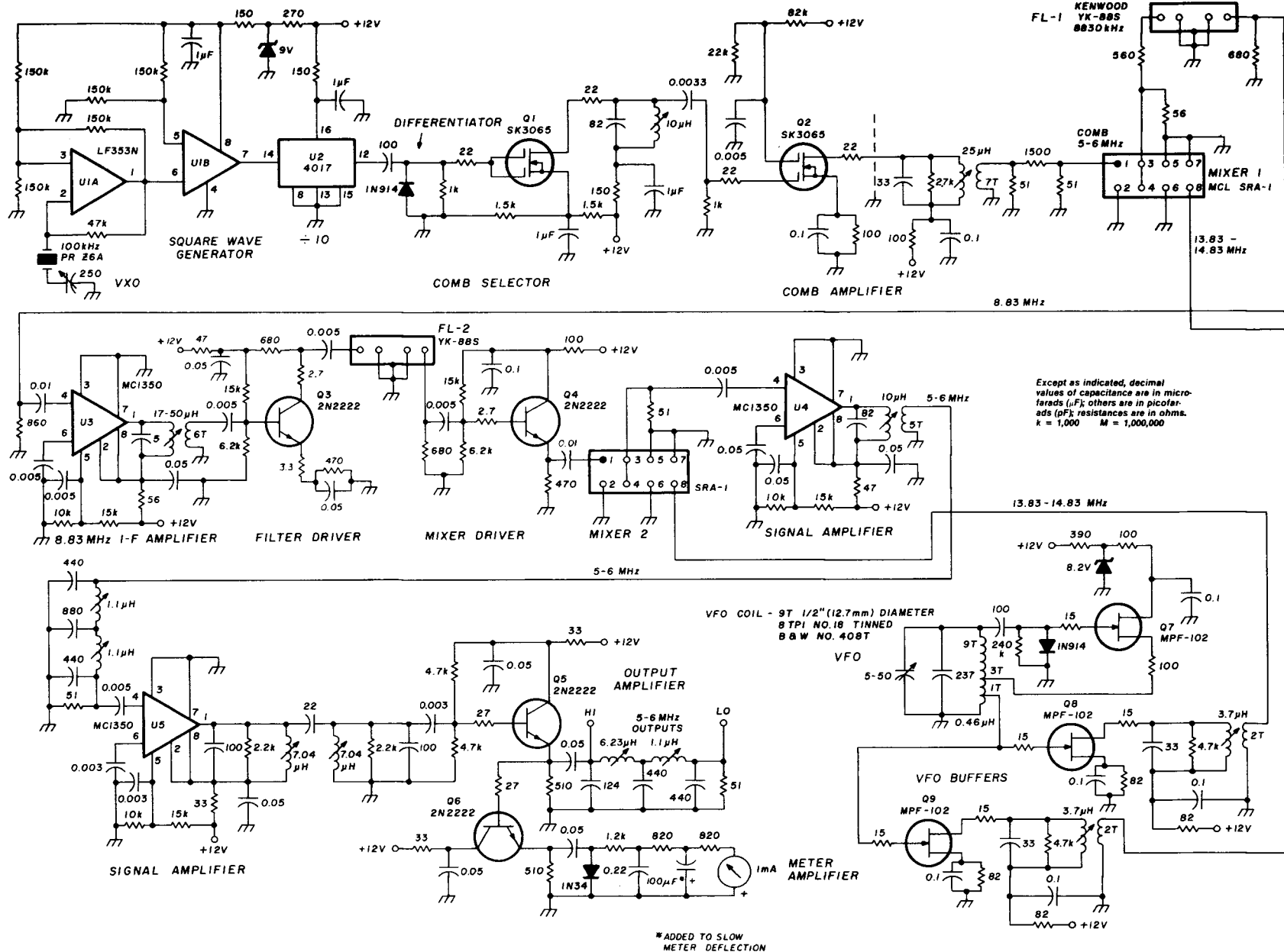


fig. 3. Detailed circuit diagram for one form of the synthesizer. Details of the design and suggestions for improvement are contained in the text.

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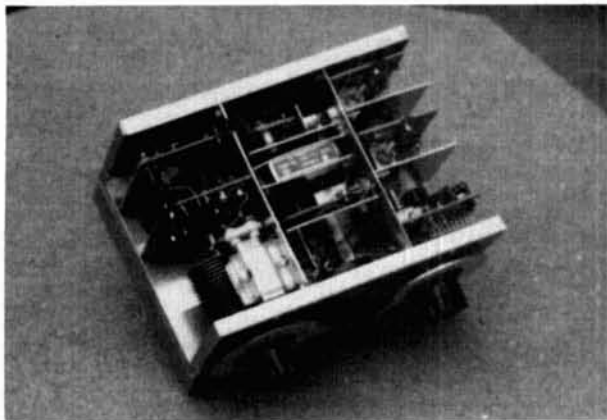
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Left, front-to-back, VXO through comb amplifier. Right, front-to-back, VFO and VFO buffers. Center, balance of circuitry through U4 output; front-to-back, Mixer No. 1, FL-1, U3, Q3, FL-2, Q4, Mixer No. 2, U4.

cans. The coils in chassis No. 1 are 3/8-inch slug-tuned ceramic unshielded forms made by Miller. A shielded cable containing power supply leads and the DC for the selection indicator interconnects the two chassis. The RF interconnection is made by means of coax cable.

Chassis No. 1 is divided into three main shielded compartments by means of two aluminum shields running front-to-back. The left compartment contains U1 through Q2. The right compartment is subdivided into three shielded sections and contains Q7, Q8, and Q9, front-to-back. The remainder of the circuit through U4 is contained in the center section. Most of the circuits are supported on perfboard placed parallel to the front panel. Push-in terminals are used where required. Care is taken to be sure that the mixer grounds are of low inductance and resistance — otherwise balance will be degraded. Similarly, the connections to the crystal filter grounds and case should be very solid, and the input and output circuits well separated and shielded; otherwise, unwanted signals may bypass the filters. The VXO tuning shaft is directly driven by a knob calibrated in kHz. The VFO shaft is driven by a 6:1 reduction drive dial calibrated in 100 kHz steps. Coax cable running below the bottom plate carries the RF interconnections between the shielded compartments. For mechanical reasons, the ends of the coax are exposed. To correct this, a shield resembling an inverted chassis is placed below the bottom plate and secured with metal spacers. This shield also raises the knobs above the table, increasing the ease of tuning.

comments

The writer's experience with low frequency VXO circuits is limited to one crystal of ancient manufacture. Before proceeding further, a prospective builder should experiment to be sure his 100 kHz crystal will pull 200 Hz. (Note that the exact fre-

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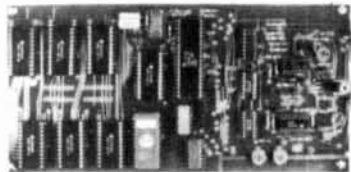
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- Model PK1 can be customized for Commercial Systems.

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quencies do not matter; the difference between the maximum and minimum frequencies must be at least 200 Hz.) If this much pull cannot be obtained, the circuit will have to be re-designed using the equations contained above.

It should be clear that this is a first model which is certainly not of optimum design. Its defense is that it presents no surprises. The second crystal filter, used to reduce noise generated in the IF amplifier and to increase the out-of-band attenuation, may not be needed. The filters associated with U5 and Q5 could be simplified and improved. The crystal could be placed in an oven for better stability. Finally, the VFO reduction drive should have a higher ratio, (preferably 100:1, to ease the tuning); if not, some form of electrical bandspreeding could be used.

acknowledgements

I appreciate the interest in this project expressed by many fellow hams. In particular, I wish to express my gratitude to Dick Schellenbach, W1JF, and Skip Marsh, W6TFQ, for their help and encouragement.

references

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ham radio

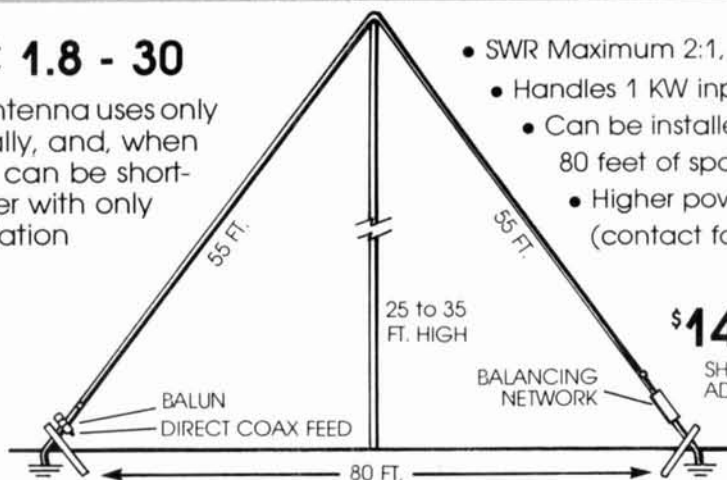
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elliptic lowpass audio filter design

Use standard
88 and 134 mH inductors
for deep attenuation
filter response

The elliptic lowpass filter is frequently used by the Radio Amateur for speech filtering because of its sharp selectivity, and many articles have been published on this particular application.¹⁻⁶ Unfortunately, the audio filter design procedures currently used by the Amateur are derived from engineering texts, and therefore are more suited to commercial practices. For example, the commercial filter designer develops his design based on a particular cutoff frequency and impedance level, and then orders the inductors from an inductor manufacturer. Capacitors are obtained from electronic distributors, measured, and then paralleled as required to get the design values within the specified tolerance. In Amateur applications, in which requirements are less stringent, variations in cutoff frequency and impedance level can be exchanged for more convenient component values. The Amateur also places greater importance on being able to use an inexpensive source of standard-value inductors than does the commercial designer. This is because the Amateur usually builds only one or two filters and therefore cannot take advantage of the large-volume discount prices available to the commercial designer.

In the discussion that follows, a simple design procedure for passive LC audio elliptic lowpass filters that use inexpensive 88 and 134-mH inductors is presented.

common filter types

The two most frequently used filter types in Amateur Radio practice are the Chebyshev and the elliptic. Both can be designed for any level of passband ripple; the elliptic is more versatile because it also can be designed for any level of stopband attenuation.⁷

Also, the attenuation rise of the elliptic can be made much more abrupt than the Chebyshev.

In RF harmonic filtering, attenuation greater than 40 dB one octave from the cutoff frequency is usually adequate; a 7th-degree Chebyshev design is most suitable because its response provides more than 42 dB attenuation at twice the cutoff frequency and it has a simple ladder configuration (four shunt capacitors alternating with three series inductors). However, for speech filtering applications, it is important that the filter attenuation rise as quickly as possible to 40 dB or more. Consequently, the 7th-degree elliptic design is preferred over the Chebyshev. Greater than 40 dB of attenuation occurs at a frequency only 1.2 times the cutoff frequency of an elliptic filter compared to two times the cutoff frequency of a Chebyshev design.

Fig. 1 shows the schematic diagram and the typical attenuation response of a 7th-degree elliptic filter. A family of elliptic designs is available that shows 45 to 65 dB of stopband attenuation and uses the *same value of inductance for L4 and L6*. This means that only two different inductance values are needed (L2, and L4, L6). One value can be the fixed 88 or 22-mH value of the standard surplus inductor, and the other can be obtained from a modified 134 or 33.5-mH inductor.

inductor types

The two inductor types suitable for the elliptic filter construction are shown in fig. 2. Both types (one potted) have molybdenum-permalloy cores and the optimum frequency range is from 300 Hz to 20 kHz. Both types have two windings that can be connected either in parallel or series aiding. Fig. 3 shows the inductors with instructions for connecting the windings in series or parallel.

The unpotted inductor has an inductance of 134 or 33.5 mH in the series or parallel connection, respectively. The potted inductor has values of 88 or 22 mH in the series or parallel connection. The potted induc-

By Ed Wetherhold, W3NQN, Honeywell, P.O. Box 391, Annapolis, Maryland 21404

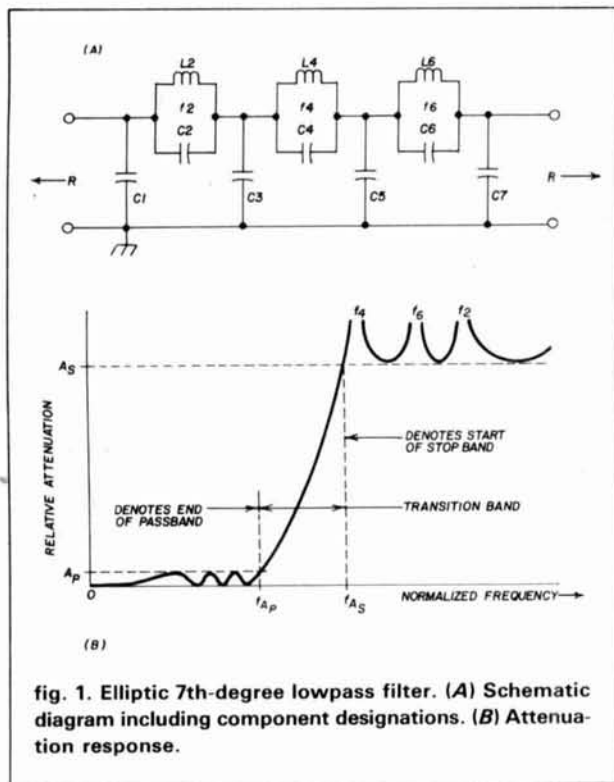


fig. 1. Elliptic 7th-degree lowpass filter. (A) Schematic diagram including component designations. (B) Attenuation response.

tor is designed for mounting on a flat surface by inserting a 1/4-inch diameter stud through a hole and then by placing a Tinnerman mounting clip over the stud to secure the inductor. The center inductor in fig. 2 is shown with a Tinnerman mounting clip.

Since turns cannot be removed from the potted inductor to vary the inductance, it must be used as is in applications requiring a 22 or 88-mH value. Because turns can be removed from the unpotted inductor, many different elliptic designs are possible, and eight different designs have been selected and calculated for stopband attenuation (A_s) levels ranging from 45.6 to 64.5 dB. The total turns needed to be removed from the unpotted inductor for these eight designs varies from eight to one hundred turns. Procedures and equations for calculating the total turns to remove for any inductance value that may be needed are given in appendix A.

7th-degree elliptic lowpass filter

The maximum amplitude of the passband ripple (A_p) (see fig. 1B) is less than 0.11 dB for these designs. The cutoff frequency is designated f_{Ap} and is defined as that frequency at which the attenuation first exceeds the A_p level. The normalized value of f_{Ap} is unity, and all other frequencies are referenced to it. The start of the stopband is denoted by f_{As} and occurs where the minimum stopband attenuation (A_s) is first achieved. The band of frequencies between the end of the passband and the start of the stopband is

called the transition band. It is desirable to minimize the transition band for best filter selectivity. An indication of the degree of selectivity is the f_{As}/f_{Ap} ratio — the smaller the ratio, the more selective the filter. Since the normalized value of f_{Ap} is unity, the normalized value of f_{As} indicates filter selectivity. The f_{As} values in the eight designs vary from 1.2 to 1.37. Note: selectivity and higher stopband attenuation are tradeoffs.

In audio filtering applications, a passband ripple of less than 0.2 dB (as found in the eight designs), is sufficiently low and won't be discussed further.

L2-C2, L4-C4, and L6-C6 resonant circuits produce attenuation peaks in the response at f_2 , f_4 , and f_6 (see fig. 1). If these attenuation peaks are correctly placed, the minimum stopband attenuation will theoretically never drop below the A_s value.

simplified design procedure

Table 1 lists all filter parameters: inductances, reflection coefficients, A_p , f_{As} , and A_s . The reflection coefficient, used by filter designers to categorize a specific design, is directly related to A_p . The relationship between R (filter termination impedance) and f_{Ap} is given in columns R and F of table 1. R can be calculated if f is known, or vice versa. The L2 and L4, L6 columns list two inductance values which correspond to the series or parallel connections of the inductor windings.

Table 2 lists the normalized component values and frequencies of the same eight designs in table 1. This data is used to calculate the actual component values and frequencies of the filter after the termination resistance and cutoff frequency are selected.

The designs in tables 1 and 2 are keyed to each other either by the design number or by the L2/L4

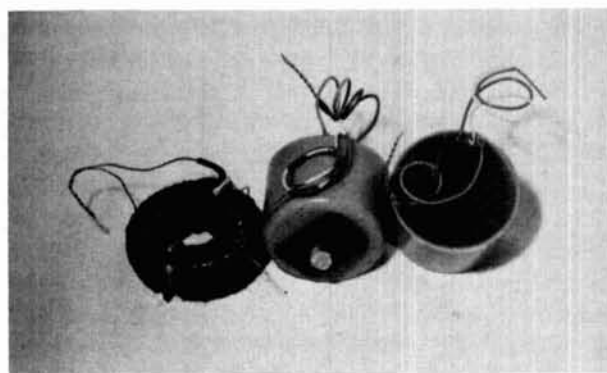
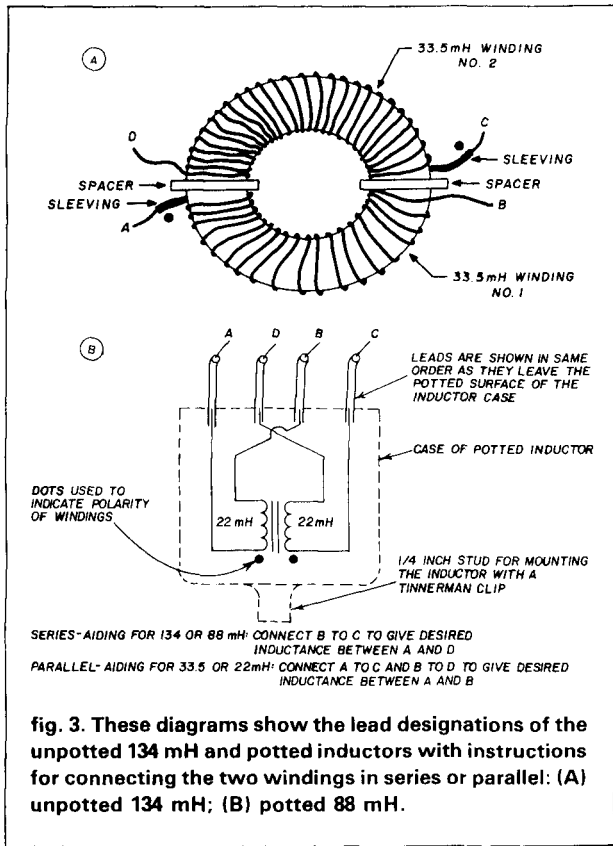


fig. 2. The two inductor types used in constructing the filter. Each has two windings which can be connected in series or parallel. The large unpotted inductor (L2) has a maximum inductance of 134 or 33.5 mH and the potted inductors (L4 and L6) have 88 or 22 mH for the series or parallel connections, respectively. All inductors are shown with their leads in the parallel connection. One of the potted inductors is shown with a Tinnerman mounting clip.



ratio. These ratios range from 1.25 to 1.50, and correspond to L_2 values from 110 to 132 mH (in the series-aiding connection) or from 27.5 to 33.0 mH (in the parallel-aiding connection). The purpose of each table will be explained before demonstrating the design procedure.

Before selecting a design from **table 1**, the cutoff frequency and the approximate impedance level must first be selected. Because the cutoff frequency is usually more critical, its value is fixed, and R is calculated using the equations under column R or by using the graph in **fig. 4**. The "A" designs are best suited for cutoff frequencies below 2 kHz, while the "B" designs are best for cutoff frequencies above 2 kHz.

In most cases, an exact match between design and circuit impedance is not possible. However, resistive matching pads (see **appendix C**) can be used if the difference between the circuit and filter impedances is less than about 30 percent. For greater impedance differences, matching transformers must be used. The disadvantage of not being able to select a specific and independent filter termination impedance is a consequence of using fixed inductor values, worthwhile because standard-value surplus inductors can be used. The filter performance parameters are next reviewed to see if they are fully satisfactory. If, for example, a minimum stopband attenuation exceeding 55 dB is re-

quired, then only designs 1 through 3 should be considered. After a design is selected from **table 1**, and normalized values of the same numbered design in **table 2** are scaled to the impedance level and cutoff frequency.

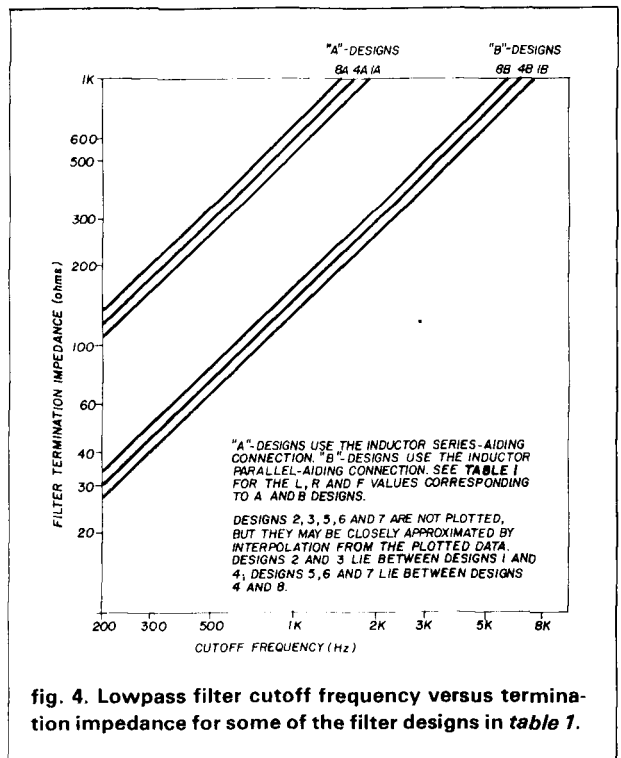
design example

To design a speech filter that has a 3 kHz cutoff frequency and a source and load impedance of 500 ohms, a typical procedure would consist of the following steps:

(a) Follow the 500-ohm line across the graph in **fig. 4** until it intersects with the slanted lines in the vicinity of 3 kHz; design 8B provides the closest match. If a lower or higher cutoff frequency is required, the filter impedance level must be changed to correspond to the new cutoff frequency. If the new filter impedance is within 30 percent of 500 ohms, resistive padding can be used to match the filter to the 500-ohm system. For greater impedance differences, suitable matching transformers must be used to prevent excessive signal amplitude loss (see note 1).

(b) Refer to **table 1** to check the performance parameters of design 8B. If f_{A_s} and A_s are satisfactory, then design 8B can be used. For design 8B, $L_4 = L_6 = 22 \text{ mH}$, $L_2 = 33 \text{ mH}$, $L_2/L_4 = 1.500$, and $R.C. = 8.06 \text{ percent}$.

(c) Find the exact f_{A_p} cutoff frequency for $R = 500$



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ohms from the equation in column f of table 1: $f = 6.0137(500) = 3007 \text{ Hz}$. Use 3007 Hz as the cutoff frequency and $R = 500 \text{ ohms}$ in the following calculations. (By a fortunate coincidence, in this particular case both the f and R values are virtually identical with the desired values. This will not always be the case, and it is up to the ingenuity of the filter designer to find an optimum combination of parameters that best satisfies the design requirements. This is one of the things that makes filter designing a challenge.)

(d) Calculate the capacitance and inductance scaling factors, C_s and L_s where R and f are in ohms and Hertz:

$$C_s = 1/(R \cdot f) = 1/(500)(3007) = 0.6651 \cdot 10^{-6}$$

$$L_s = R/f = 500/3007 = 166.3 \cdot 10^{-3}$$

(e) Calculate the L and C component values by multiplying the normalized component values of design 8, table 2, by the C_s and L_s scaling factors. Frequencies are obtained by multiplying the normalized frequencies of design 8, table 2, by the cutoff frequency, 3007 Hz. The component and frequency values obtained are:

$$C1 = 0.1289f (0.6651 \cdot 10^{-6}) = 0.0857 \mu\text{F}$$

$$C2 = 0.02561f (0.6651 \cdot 10^{-6}) = 0.01703 \mu\text{F}$$

$$C3 = 0.2084f (0.6651 \cdot 10^{-6}) = 0.1386 \mu\text{F}$$

$$C4 = 0.1291f (0.6651 \cdot 10^{-6}) = 0.0859 \mu\text{F}$$

$$C5 = 0.1819f (0.6651 \cdot 10^{-6}) = 0.1210 \mu\text{F}$$

$$C6 = 0.09836f (0.6651 \cdot 10^{-6}) = 0.0654 \mu\text{F}$$

$$C7 = 0.08079f (0.6651 \cdot 10^{-6}) = 0.0537 \mu\text{F}$$

$$L2 = 0.19845H (166.3 \cdot 10^{-3}) = 33.0 \text{ mH}$$

$$L4,6 = 0.1323H (166.3 \cdot 10^{-3}) = 22.0 \text{ mH}$$

$$f_{A_p} = 1.000 (3007 \text{ Hz}) = 3007 \text{ Hz}$$

table 1. Lowpass filters: inductor values and L2/L4 ratios with related parameters for selected 7th-degree elliptic designs where $L4 = 1.6$. [For filter construction using two potted 88-mH inductors and one modified 134-mH inductor.]

design no.	L2/L4 ratio	L2	L4,6	R.C. (%)	A _p (dB)	f _{A_s}	A _s (dB)	R (ohms)	f (Hz)	
		(mH)								
1	A	1.250	110.0	88	15.36	0.1037	1.3703	64.50	0.5279(f)	1.894(R)
	B		27.5	22					0.1320(f)	7.577(R)
2	A	1.300	114.4	88	12.80	0.07171	1.3166	59.16	0.5515(f)	1.8133(R)
	B		28.6	22					0.1379(f)	7.2532(R)
3	A	1.333	117.3	88	11.56	0.05842	1.2888	56.19	0.5689(f)	1.7578(R)
	B		29.3	22					0.1422(f)	7.0313(R)
4	A	1.380	121.4	88	10.23	0.04571	1.2573	52.61	0.5947(f)	1.6817(R)
	B		30.4	22					0.1487(f)	6.727 (R)
5	A	1.400	123.2	88	9.77	0.04164	1.2458	51.25	0.6060(f)	1.6501(R)
	B		30.8	22					0.1515(f)	6.6005(R)
6	A	1.440	126.7	88	8.98	0.03517	1.2258	48.77	0.6293(f)	1.5890(R)
	B		31.7	22					0.1573(f)	6.356 (R)
7	A	1.460	128.5	88	8.64	0.03257	1.2169	47.64	0.6411(f)	1.5597(R)
	B		32.1	22					0.1603(f)	6.239 (R)
8	A	1.500	132.0	88	8.06	0.02829	1.2012	45.56	0.6652(f)	1.5034(R)
	B		33.0	22					0.1663(f)	6.0137(R)

Notes.

1. The unpotted inductor (L2) has two 33.5 mH windings that can be connected in series or parallel aiding to give 134 or 33.5 mH, respectively. The potted inductors (L4 and L6) have two 22 mH windings which can be connected in series or parallel aiding to give 88 or 22 mH. Fig. 3 shows the series and parallel aiding connections.

2. See fig. 1A for the schematic diagram and fig. 1B for the typical attenuation response of the above filter designs. See table 2 for the normalized component values and frequencies.

3. To obtain the L2 inductance values, remove an equal number of turns from both windings of the unpotted inductor in accordance with the following equations (as shown in appendix A):

(a) for all "A" designs, $T_s = 532 - 45.957\sqrt{L_s}$, where T_s is the number of turns removed from each winding (total turns removed = $2 \cdot T_s$), and L_p = the desired L2 inductance in the series-aiding connection.

(b) for all "B" designs, $T_p = 532 - 91.914\sqrt{L_p}$, where T_p is the number of turns removed from each winding (total turns removed = $2 \cdot T_p$), and L_s = the desired L2 inductance in the parallel-aiding connection.

For example, for L_s or $L_p = 132$ or 33 mH , remove four turns from each winding and connect the windings in the series or parallel-aiding connection, respectively.

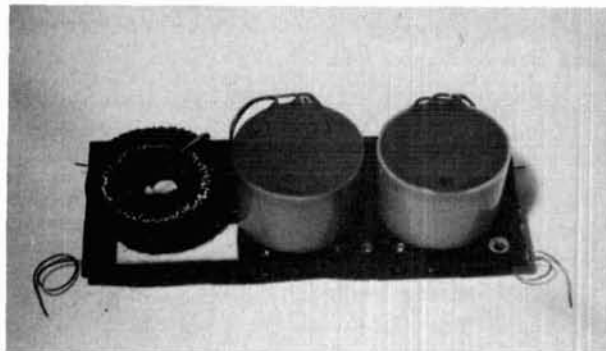
$$\begin{aligned}
 f_{A_s} &= 1.201 (3007 \text{ Hz}) = 3611 \text{ Hz} \\
 f_2 &= 2.233 (3007 \text{ Hz}) = 6715 \text{ Hz} \\
 f_4 &= 1.218 (3007 \text{ Hz}) = 3663 \text{ Hz} \\
 f_6 &= 1.395 (3007 \text{ Hz}) = 4195 \text{ Hz} \\
 \text{with } R &= 500 \text{ ohms and } A_s = 45.6 \text{ dB}
 \end{aligned}$$

As a check on the correctness of the values of (C2, L2), (C4, L4) and (C6, L6), the resonant frequencies of these parallel tuned circuits can be calculated and should agree to within about 0.1 percent with the values of f_2 , f_4 , and f_6 calculated from the scaling procedure.

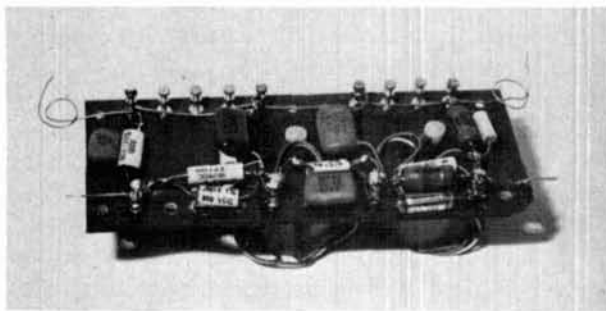
filter construction

Figs. 5A and 5B are photographs of the completed filter design 8B. The inductors and capacitors are mounted on opposite sides of a 2 x 5-inch terminal board. The unpotted inductor is fastened to the terminal board with 3M Scotch™ tape, and the two potted inductors are mounted with their 1/4-inch mounting studs and Tinnerman clips.

The tolerances of the shunt capacitors (C1, 3, 5, and 7) are not critical, and values within 5 percent of the design values are satisfactory. Two capacitors can be paralleled when necessary to obtain the required value. The calculated values of the resonating capacitors (C2, 4, and 6) should be used only as a guide. This is because the actual resonant frequencies of the three tuned circuits are of primary importance, while the exact values of capacitance are of secondary importance. The 88-mH value of the potted inductors is only nominal, and it can vary as low as 86 mH. It seldom is greater than 88 mH. The winding capacities of the potted inductors in the 88- and 22-mH connections are about 68 and 176 pF, respectively. The winding capa-



(A)



(B)

fig. 5. The photographs of the completed filter show how the inductors and capacitors are mounted on opposite sides of a standard terminal board. The filter assembly can then be mounted in a standard 5 x 4 x 3-inch aluminum mini-box. In (A), the unpotted inductor (L2) is on the left followed by L4 and L6. In (B), capacitor C1 is on the left. The common ground return wiring is at the top of the terminal board.

cities of the unpotted inductor in the 134- and 33.5-mH connections are about 68 and 148 pF, respectively. The simplest procedure for tuning the three resonant circuits of the elliptic filter is to assemble all the filter com-

table 2. Lowpass filter: normalized component values and frequencies of selected 7th-degree elliptic designs where $L4 = L6$. (Values are normalized for an f_{A_p} cutoff frequency of 1 Hz and 1-ohm terminations.)

design no.	L2/L4 ratio	C 1	C 2	L 2	C 3	C 4	L4&L6	C 5	C 6	C 7	f_{A_s}	f_2	f_4	f_6
		(Farad)			(Henry)	(Farad)		(Henry)	(Farad)					
1	1.250	0.1760	0.01627	0.2083	0.2671	0.07819	0.1667	0.2439	0.05647	0.1438	1.370	2.734	1.394	1.640
2	1.300	0.1611	0.01832	0.2074	0.2495	0.08863	0.1596	0.2251	0.06487	0.1254	1.317	2.582	1.338	1.564
3	1.333	0.1535	0.01964	0.2063	0.2402	0.09552	0.1547	0.2152	0.07047	0.1155	1.289	2.501	1.309	1.524
4	1.380	0.1448	0.02141	0.2042	0.2293	0.1051	0.1480	0.2036	0.07830	0.1038	1.257	2.407	1.276	1.478
5	1.400	0.1416	0.02214	0.2033	0.2252	0.1091	0.1452	0.1993	0.08165	0.09931	1.246	2.372	1.264	1.462
6	1.440	0.1360	0.02357	0.2014	0.2179	0.1172	0.1398	0.1916	0.08835	0.09126	1.226	2.310	1.243	1.432
7	1.460	0.1335	0.02426	0.2004	0.2145	0.1211	0.1373	0.1881	0.09169	0.08758	1.217	2.283	1.234	1.419
8	1.500	0.1289	0.02561	0.19845	0.2084	0.1291	0.1323	0.1819	0.09836	0.08079	1.201	2.233	1.218	1.395

Notes:

- Fig. 1A shows the interconnections of the capacitors and inductors. Fig. 1B shows the cutoff frequency and attenuation peaks corresponding to the f_1 , f_2 , f_4 , and f_6 values.
- The normalized data for the filter design numbers in this table also apply to similarly numbered designs in table 1.

ponents except for the resonating capacitors, C2, C4 and C6. Connect the partially assembled filter between a signal source and a resistive load, both having the design impedance of the filter. Set the signal source to the frequency of the first resonant circuit to be tuned, and monitor the filter output voltage across the load resistor with an AC VTVM. Resonate the inductor of the first resonant circuit with a calibrated capacitor decade. Resonance will be indicated by a minimum reading on the AC VTVM. Read the capacitance required for resonance from the capacitor decade, and then place this value of capacity across the inductor using as many paralleled capacitors as required. Repeat the same procedure for the other two resonant circuits. Using this procedure, it is possible to tune the resonant circuits to within 0.25 percent of the design frequency values.

For optimum stopband performance, the ground leads of C1, 3, 5, and 7 must be independent of each other to minimize any common ground impedance.

testing the filter

Fig. 6 shows the measured filter attenuation compared with the computer-calculated response. Close agreement between measured and calculated responses in the pass and transition bands indicates that a good filter design and assembly procedure were used. The lower-than-expected attenuation peaks at f_2 , f_4 , and f_6 are because of a common ground impedance in the ground leads of the four shunt capacitors. If separate ground leads are used from the shunt capacitors, the peak attenuation levels can be increased by more than 10 dB. This will cause the measured minimum attenuation A_s levels to increase slightly.

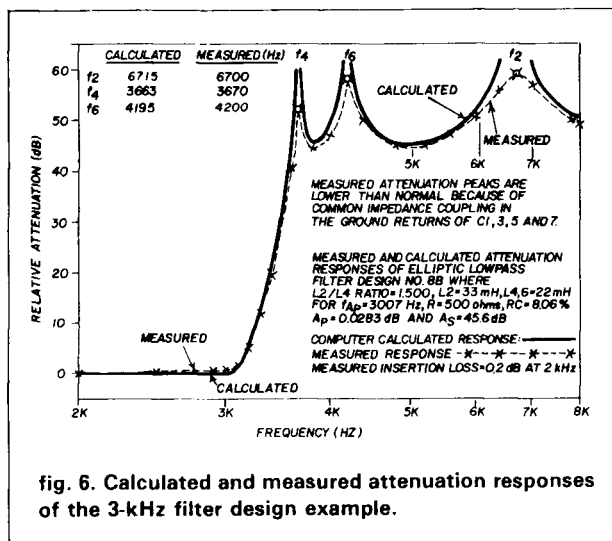


fig. 6. Calculated and measured attenuation responses of the 3-kHz filter design example.

The passband insertion loss is only 0.2 dB. This low value is due to the relatively high- Q inductors in the 3 kHz to 8 kHz frequency range. For much lower cutoff frequencies, the inductor Q 's will be lower and the insertion loss will be higher.

obtaining surplus inductors

The potted 88-mH inductors have been provided to the author at no charge by the C & P Telephone Company of Maryland for distribution to those Radio Amateurs wishing to build audio filters. The 134-mH inductors were purchased by the author and are being offered at cost to those wishing to build the filter designs described in this article. Your request for inductors should be addressed to me and accompanied by a description of your filtering application and a 4 x 9-1/4 inch SASE.

summary

Until now, the Radio Amateur has been hindered in using high-performance audio lowpass filters due to the lack of a simplified design procedure and a source of high- Q and inexpensive inductors. This article explains how these problems have been solved. By using computer-aided calculations, designs from a special family of elliptic lowpass filters have been derived that can use two surplus potted 88-mH inductors and one unpotted 134-mH inductor.

Data was tabulated for eight 7th-degree elliptic lowpass designs suitable for the audio frequency range, and a simplified design procedure was demonstrated by designing and building a 3-kHz, 500-ohm lowpass filter. Comparisons were made between the measured and computer-calculated responses to show that the design and construction procedures were valid. It was explained how these design procedures usually result in non-standard impedance levels, and how this inconvenience can be alleviated by using padding resistors or inexpensive matching transformers. A source of surplus inductors for the filter construction was given, and a procedure was explained for deriving an equation for modifying the unpotted inductor for a specific inductance.

acknowledgement

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appendix A

For a series-aiding connection calculate the total number of turns, T_o , as follows:

(a) Measure the original inductance, L_o , with the windings connected in series-aiding (SA) using an inductance bridge. An indirect method of inductance measurement is to resonate the inductor with a precisely known capacitance (± 0.5 percent) while measuring the resonant frequency with a digital frequency counter.⁹ Calculate the inductance from the equation:

$$L_{(mH)} = 25.33/(f^2C) \quad (1)$$

where f and C are in kHz and μF .

Use coupling capacitors of approximately 500 pF each to isolate the tuned circuit from the signal generator and the AC VTVM.

(b) Remove 50 turns from each winding and connect the windings in series aiding. Measure the modified inductance, L_r .

(c) Calculate $T_o = 100 \cdot R / (R - 1)$ (2)

where $R = \sqrt{L_o/L_r}$
and T_o = original number of turns on the unmodified inductor
 L_o = original inductance (mH) in the series-aiding (SA) connection
 L_r = inductance (mH) for the SA connection after 100 turns were removed

For example, if $L_o = 134 \text{ mH}$ and $L_r = 110 \text{ mH}$ for 50 turns removed from each winding, then $R = 1.10371274$ and $T_o = 1064$.

The unique value of T_o is used to find the equation that gives the total number of turns to be removed from any inductor (of the same type) in order to obtain a specific inductance.

$$T_s = 0.5 (T_o \cdot S \sqrt{L_s}) \quad (3A)$$

where $S = (T_o - T_r) / \sqrt{L_r}$ (3B)

and T_s = number of turns to remove from each winding to obtain L_s
 L_s = desired inductance (mH)
 T_o = previously determined number of turns on the unmodified inductor
 T_r = total turns removed to get L_r
 L_r = inductance (mH) when T_r turns are removed

Example: a 134-mH inductor has a $T_o = 1064$, with $L_r = 110$ for $T_r = 100$

$$S = (1064 - 100) / \sqrt{110} = 91.914$$

$$\begin{aligned} \text{then } T_s &= 0.5 (1064 \cdot 91.914 \sqrt{L_s}) \\ &= 532 \cdot 45.957 \sqrt{L_s} \end{aligned}$$

For example, find T_s to get 132 mH from the 134-mH inductor.

$$T_s = 532 \cdot 45.957 \sqrt{132} = 532 \cdot 528.0$$

$$\begin{aligned} &= 4 \text{ turns removed from each winding.} \\ \text{Total turns removed} &= 8 \end{aligned}$$

For a parallel-aiding connection:

$$T_p = T_o/2 \cdot S \sqrt{L_p} \quad (4)$$

where T_o and S were previously defined, and
 T_p = number of turns to remove from each winding to obtain L_p
 L_p = desired inductance (mH)
 $T_p = 91.914 \sqrt{L_p}$ (5)

Find T_p in order to obtain 33 mH from the 134-mH inductor with its windings in the parallel-aiding connection:

$$\begin{aligned} T_p &= 91.914 \sqrt{33} = 532 \cdot 528.0 \\ &= 4 \text{ turns removed from each winding} \\ \text{Total turns removed} &= 8 \end{aligned}$$

appendix B

Equations relating reflection coefficient and A_p :

$$R.C. (\text{percent}) = 100 \sqrt{1 - 0.1^x} \quad (1)$$

where $x = A_p/10$ and A_p is the maximum passband ripple amplitude in dB.

$$A_p (\text{dB}) = -10 \cdot \log_{10}(1 - \rho^2) \quad (2)$$

where $\rho = \text{percent } R.C. / 100$

For example, if $R.C. = 8.058 \text{ percent}$, then $A_p = 0.0282913 \text{ dB}$.

appendix C

determination of filter matching pad resistances

When the source impedance (Z_s) or load impedance (Z_L) is within 30 percent of the termination impedance required by the filter, a single resistor can be placed in series or parallel with Z_s or Z_L to produce the termination impedance required by the filter. The slight loss in signal level can usually be compensated for by simply increasing the volume control on your receiver. For greater impedance differences, a suitable matching transformer must be used (see note 1).

(A) When Z_s or Z_L is less than the required filter termination impedance place a resistance in series with the Z_s or Z_L so that the sum of the two equals the desired filter termination impedance.

(B) When Z_s or Z_L is greater than the required filter termination impedance place a resistance in parallel with the Z_s or Z_L so that the parallel combination equals the desired filter termination impedance. If the added parallel resistance is " R_p ," then:

$$R_p = (S \cdot F) / (S - F)$$

where S = source impedance and
 F = filter impedance with all values in ohms.

For example, if a 500-ohm filter is to be matched to a 350-ohm source and a 600-ohm load, then place a 150-ohm resistor in series with the 350-ohm source, and connect a 3 kilohm resistor in parallel with the 600-ohm load.

Note 1: A large selection of inexpensive audio transformers of different impedances at power levels of 0.075, 0.2, 0.4, and 2-watts output can be obtained from Mouser Electronics, 11433 Woodside Avenue, Santee, California 92071 (714-449-2222). Prices vary from 79 cents to \$1.20 for the 0.075 to 0.4-watt levels and \$2.30 for the 2-watt level.

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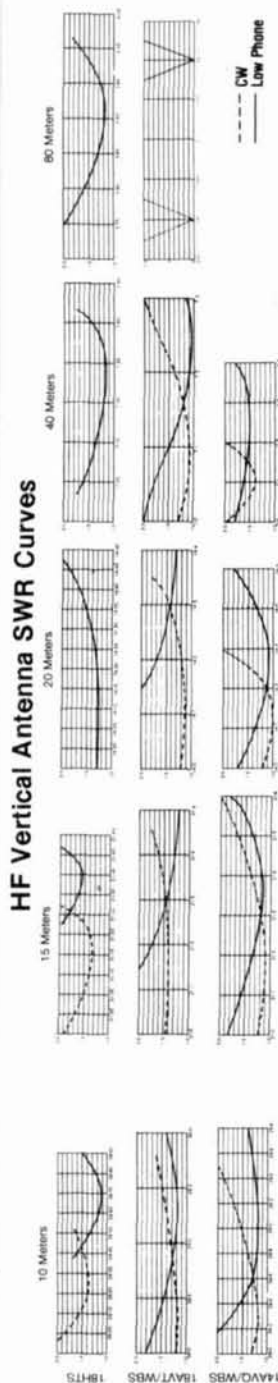
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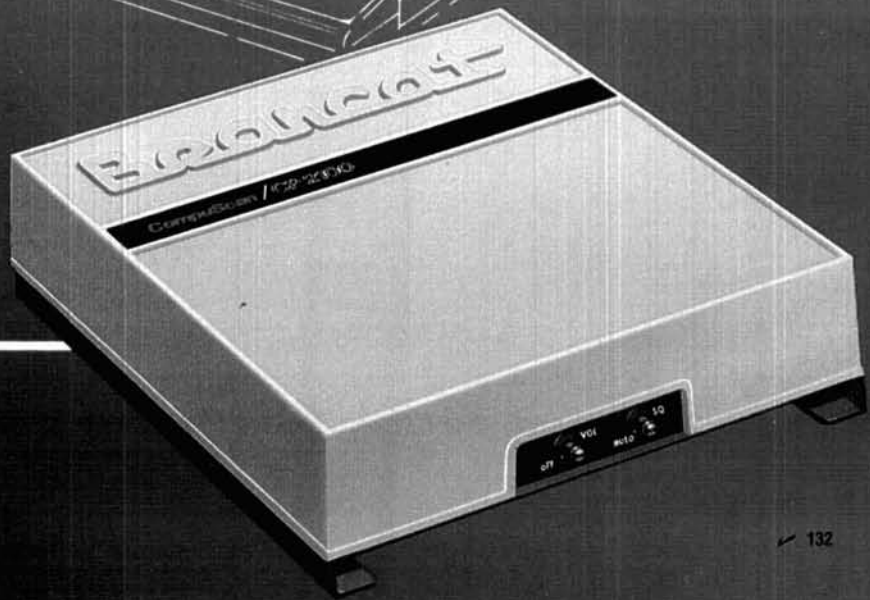
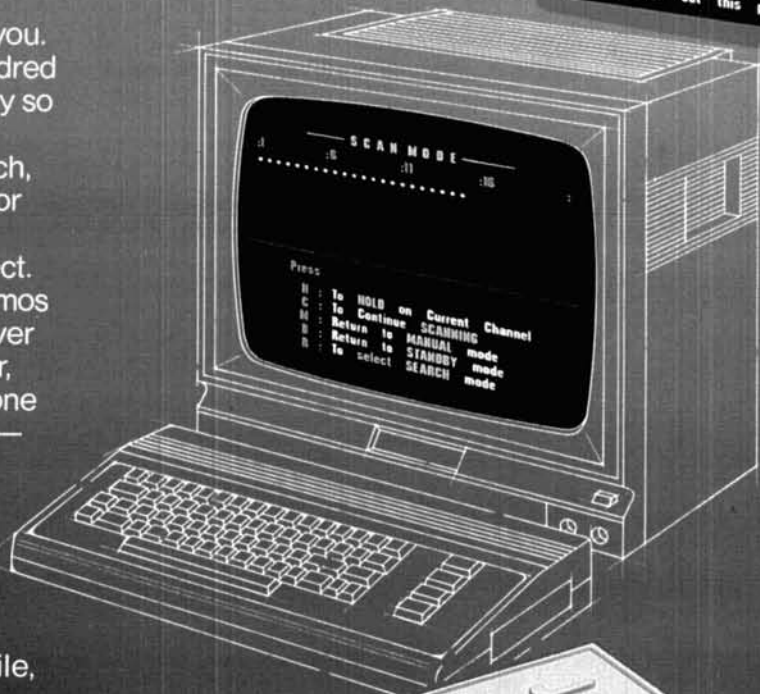
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high frequency receiver performance

Understanding and specifying good receiver operation

Receiver performance and the means for its specification is a complex and frequently misunderstood topic, for it involves many different and often unrelated parameters, all of which are important to the overall performance of the receiver. (For example, it's no good having exceptional sensitivity if the receiver drifts excessively.) When specifying performance, it is important that all parameters be specified accurately and completely. Many manufacturers and reviewers have been guilty of specifying performance inadequately, with ambiguous and misleading results. For example, the frequently quoted "sensitivity 0.5 μV " is meaningless, and "0.5 μV for 10 dB" is not much better.

It is interesting to note that the importance of some performance parameters (e.g., intermodulation and reciprocal mixing) have only recently been fully appreciated and understood. Also certain receiver design changes have actually been responsible for *reducing* performance: for example, the direct replacement of tubes by bipolar transistors in the late 1960's, and the current widespread use of frequency

synthesis, which can introduce spurious receiver responses if not designed with special care.

noise and sensitivity

One of the fundamental concepts underlying receiver performance is that of noise. So before we examine performance parameters — what they mean, how they are specified, and how they can be improved — we'll take a brief look at the physics of noise.

Thermal noise is due to the random movement of particles in the effective impedance at the input to the receiver at ambient temperature. It cannot be avoided (except by cooling the whole antenna system down to near absolute zero) and is even present if a shielded 50 or 75 ohm resistor representing the antenna impedance is plugged into the receiver input. It is given by:

$$V = \sqrt{4kTBR} \quad (1)$$

where: V = RMS noise voltage
 k = Boltzmann's constant
 T = temperature in degrees Kelvin
 B = receiver bandwidth in Hz
 R = resistance/impedance in ohms

For a given antenna impedance and a nominal ambient temperature (taken as 300 degrees K or 27

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degrees C), the only way to reduce the thermal noise is to reduce the receiver bandwidth. For a 50 ohm impedance, ambient temperature of 300 degrees K, and a bandwidth of 3 kHz the above expression works out as:

$$V = \sqrt{(4)(1.38 \times 10^{-23})(300)(3000)(50)}$$

$$= 0.05 \mu V \text{ (EMF)} = -26 \text{ dB}\mu V$$

where a dB μ V is a dB relative to 1 microvolt (EMF).

Sensitivity is expressed as the smallest signal required to give a specific signal-to-noise ratio (S/N),* in a particular receiver bandwidth, and in the case of AM, for a given modulation level. In the days of noisy tubes, good sensitivity was hard to achieve without compromising other performance aspects. Now with bipolar transistors and FETs, sensitivities of 0.5 μ V for a 10 dB S/N ratio in a 3 kHz bandwidth, can easily be achieved on HF for an SSB or CW signal. As AM is usually specified at 30 percent modulation level, the figure will be 10 dB (3.16 times) worse than the SSB/CW figure; in this case, 1.6 μ V (EMF).

Noise factor (NF) can be defined as the ratio of the S/N of a hypothetically perfect (noiseless) receiver to that of a real receiver, which adds its own noise to the thermal noise. Since it is the ratio of two ratios it is independent of bandwidth, temperature, and impedance.

Typical HF spectrum noise voltages within a 3 kHz bandwidth are shown in **fig. 1**. We see at $-26 \text{ dB}\mu V$ the thermal threshold noise, and also the receiver noise for a NF of 10 dB, at $-16 \text{ dB}\mu V$. To achieve a S/N of 10 dB a signal will need to be at $-6 \text{ dB}\mu V$ or 0.5 μ V, which means that an NF of 10 dB is equivalent to a sensitivity of approximately 0.5 μ V for a 10 dB S/N in a 3 kHz bandwidth.

However, the most significant conclusions to be drawn from **fig. 1** are concerned with atmospheric noise. This is plotted on **fig. 1** for a quiet area at a

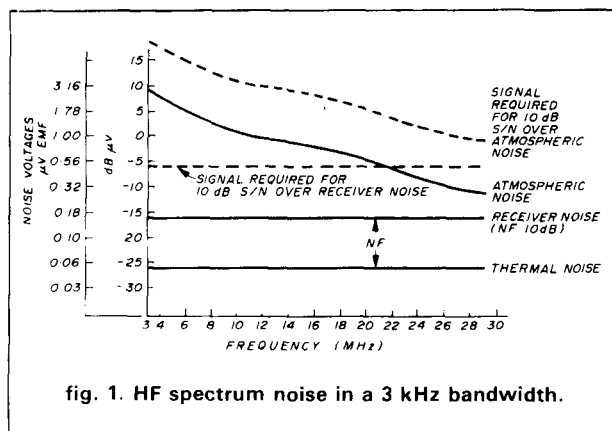


fig. 1. HF spectrum noise in a 3 kHz bandwidth.

*In fact it's more appropriate to quote signal + noise to noise (S + N/N) or even signal + noise + distortion to noise + distortion (SINAD). Note for an S/N of 10 dB or more there is little difference between the three terms.

quiet time, and is between 5 and 25 dB above receiver noise; consequently for a receiver with a noise figure of 10 dB it is the atmospheric noise, not receiver noise, that limits receiver performance. Indeed the NF could be increased to 15 dB (1 μ V EMF sensitivity) without loss of performance, except perhaps at 25 to 30 MHz. There is, therefore, no point in reducing the NF below 10 dB — especially as sensitivity can only be obtained at the expense of dynamic effects such as intermodulation. It is also worth noting that claims of 0.15 μ V (EMF) for 10 dB S/N (seen recently for an SSB transceiver) are quite impossible. Even a perfect receiver with a 0 dB NF needs 0.16 μ V ($-16 \text{ dB}\mu V$) to achieve 10 dB S/N due to the thermal threshold of $-26 \text{ dB}\mu V$.

From **fig. 1** it can be seen that under real operating conditions a receiver with a sensitivity of 0.5 μ V will in fact need between 1 μ V (at 30 MHz) and 10 μ V (3 MHz) for a 10 dB S/N ratio on HF — and this for a quiet atmosphere (and no QRM)!

The above discussion considered noise voltage in a 3 kHz receiver bandwidth. However, noise is proportional to the square root of the bandwidth. Thus if bandwidth is reduced from 3 kHz to 300 Hz, all noise voltages (thermal, receiver, and atmospheric) drop by a factor of $\sqrt{10}$ or 3.16 times (10 dB). This explains the continuing use of CW in the HF bands; a CW signal can still be copied when SSB would be lost in the noise. In fact, some operators can copy a CW signal with a S/N of around 0 dB, so the advantage over SSB can be as much as 20 dB[†]. Thus sensitivity for the same (10 dB NF) receiver can be quoted as 0.05 μ V for 0 dB S/N in a 300 Hz bandwidth.

Note that all voltages above are EMF (Electromotive Force). Recently it has become common to specify sensitivities and other parameters using potential difference (PD) instead. This practice makes sensitivity figures look twice as good (6 dB better) because, in a matched impedance system, PD is *always* half EMF (see Appendix). Care must therefore be exercised when interpreting receiver specifications; often the distinction is not made clear, usually indicating that PD is intended (see **dynamic range** for explanation).

selectivity

Selectivity used to be achieved by means of distributed tuned circuits in the IF strip, and to obtain good selectivity a low second IF was required (for example, 455 kHz). Selectivity is now usually obtained by means of crystal, mechanical, or ceramic block filters. The ideal filter response has a flat top with low ripple, and steep sides going down to a -70 dB (or

†If one considers that the human ear consists of a contiguous series of extremely narrow high Q filters (less than 30 Hz) then perhaps the ear/brain combination always works with a positive S/N ratio for intelligibility. Editor

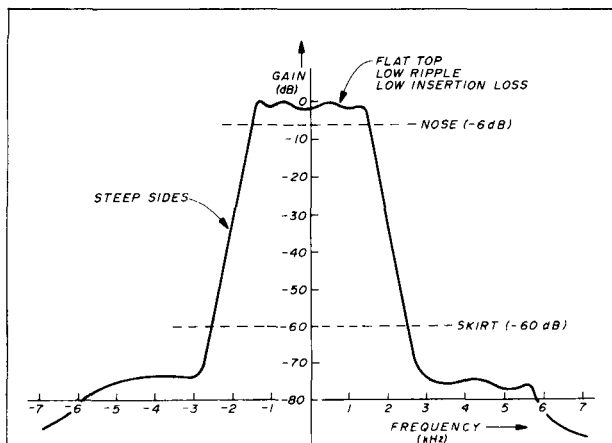


fig. 2. A desirable filter response would include a fast roll-off, low passband ripple and insertion loss, and a high stopband rejection.

greater) stopband, and extending a long way out (see fig. 2).

The old constraint of a low second IF no longer applies. In fact, it is easier to design crystal filters for higher frequencies; IFs of 1.4, 1.6, 9 and 10.7 MHz often being used.

Selectivity is usually quoted at the nose bandwidth (6 dB down), and the skirt bandwidth at 60 dB down. Good values for an 8 pole SSB filter are 2.7 kHz and 4.4 kHz, respectively. One measure of filter performance often quoted is the **Shape Factor (SF)**, which is the ratio of the skirt bandwidth to the nose bandwidth. The ideal SF is 1:1, and anything less than 2:1 for a 3 kHz SSB filter is considered good.

Mechanical and crystal filters can get quite close to the ideal, and some less expensive ceramic filters give surprisingly good results. Impedance matching into and out of a filter is of great importance and insertion loss (the loss caused by the filter in the middle of the passband) must be made up by amplification (usually less than 10 dB). A typical set of filters for a high grade communications receiver might be 8 kHz (AM), 2.7 kHz (SSB) — often with two asymmetrical filters, one for USB one for LSB — and 1.0 kHz, 300 Hz, and 100 Hz for CW. The trend in Amateur equipment is for the tightest possible SSB filter (2.4 kHz), and often 600 Hz or 300 Hz for CW are used.

The above refers to what could be called the "static" selectivity of the receiver, or the selectivity to a single signal only. For a discussion on dynamic selectivity see the section on **reciprocal mixing**.

As an interesting aside, consider the use of the audio CW filter, often used by the Amateur fraternity in lieu of a good CW filter at the IF. (The best filter of all, of course, is the human brain; a good operator can pick out and copy a weak CW signal in company

with numerous other signals because of the difference in tone. Many experienced CW operators prefer to listen in a wide bandwidth and do their own filtering even when sharp CW filters are available.) However, the audio image frequency (see fig. 3) will also give an output of exactly the same tone, and even the "human filter" will find it impossible to differentiate. But this is exactly the frequency that the audio filter cannot differentiate, either! Also, unless AGC voltage is at least partially audio-derived, strong unwanted signals in the IF passband will reduce the IF gain, hence reducing post-filter dynamic range.

Nevertheless a good multi-pole audio filter can be beneficial, especially if a linear detector is being used. (A product detector is linear; an envelope detector is not.) Also, if a steep-sided SSB filter is available, the audio image can be rejected if the BFO injection is made to coincide with the edge of the passband.

Image (second channel) rejection. In the normal superheterodyning process, a wanted signal (F_S) beats in the mixer with the local oscillator (or synthesizer output) frequency (F_{LO}), and one of the resultant products of the mixing process, usually

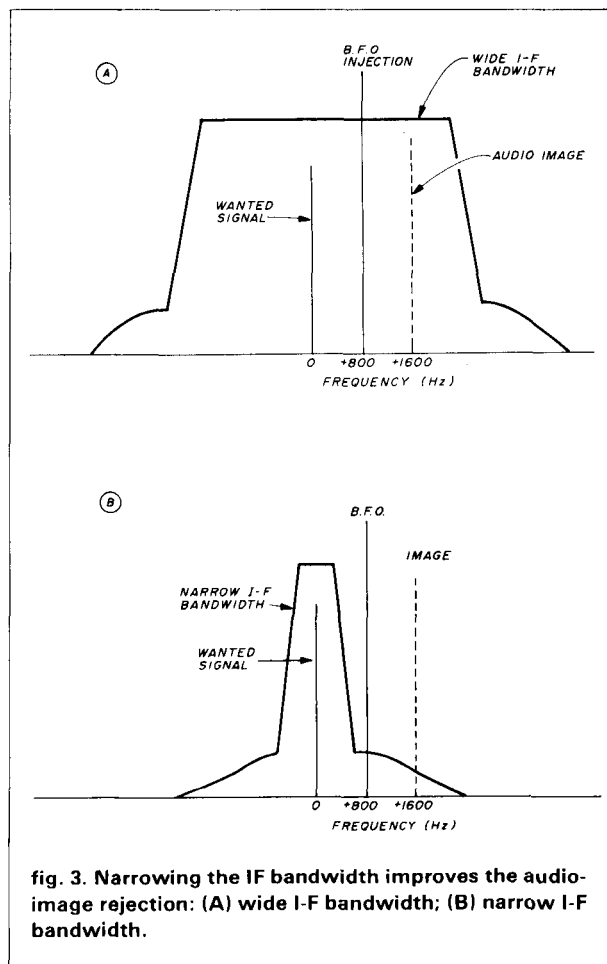


fig. 3. Narrowing the IF bandwidth improves the audio-image rejection: (A) wide I-F bandwidth; (B) narrow I-F bandwidth.

$F_{LO} - F_S$, at the intermediate frequency (IF), is passed by the IF selectivity filter.

However, another frequency called the image or second channel frequency ($F_{LO} + F_{IF}$) also beats with the local oscillator frequency to produce a product at the IF. This frequency must be rejected by some form of RF tuning, either ganged to the "tune" control or using a separate "pre-select" control; or by means of switched bandpass filters, usually automatically switched on synthesized receivers.

As the image frequency is equal to F_S plus twice the IF, the higher the first IF, the further away from F_S will be the image frequency, and the easier it will be to reject. If up-conversion techniques are used, the first IF will be in the range 40 to 90 MHz and the image frequency will also be at VHF and thus can be rejected by a simple 35 MHz low-pass filter at the receiver input.³

Image frequency rejection is specified as the ratio in dB of an unwanted signal above 1 μV to give the same output as the received 1 μV signal. 50 dB of rejection is considered poor, while 80 dB or more is desirable.

IF rejection

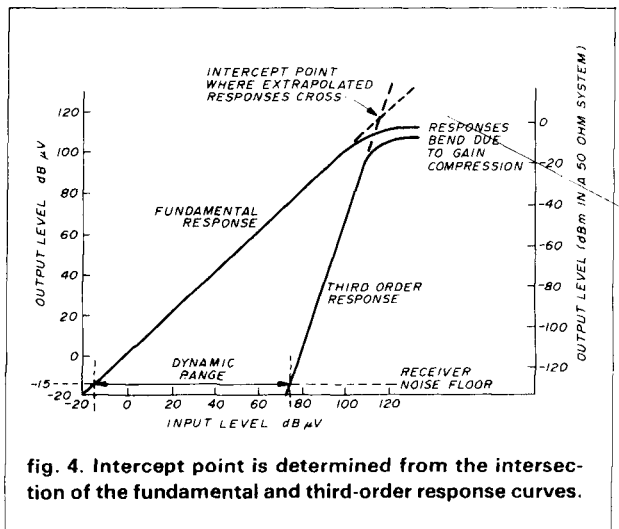
IF rejection occurs when a strong signal at a receiver IF directly breaks through the early stages and into the IF. It is specified in the same manner as image rejection with 80 dB being a desirable number. It should be quoted for all the IFs in a receiver. Often in a double conversion receiver the figure for the second IF is worse than that for the first IF. Good screening is necessary between IF and RF stages, and IF traps in early pre-IF stages can be employed to reduce IF breakthrough. Taken together, IF and image rejection are sometimes referred to as "rejection to external spuri."

dynamic effects

Dynamic interference effects such as intermodulation and cross-modulation have often been largely ignored in the past, and only in the last ten years or so has their true importance been understood. It has to be said that the replacement of tubes by bipolar transistors in the 1960s made the situation worse.* In general, dynamic effects are caused by large off-tune (off frequency) signals that cause the receiver to operate in a non-linear manner. It is these effects (together with reciprocal mixing) that currently determine the performance of the communications receiver rather than the traditional parameters of sensitivity, selectivity, and stability.

dynamic range and intercept point

Dynamic range can be loosely described as the



range of input signals over which dynamic interference effects produce outputs which are not significant (i.e., which are below the noise floor). In order to arrive at a suitable definition for dynamic range, consider the intercept point as shown in fig. 4. This occurs because the power level of the dynamic product increases at a greater rate than that for the wanted signal. Second order products increase as the square of the input (twice as many dB), and third order products as the cube (3 times in dB). (Third and second order products will be dealt with in greater detail when discussing intermodulation.)

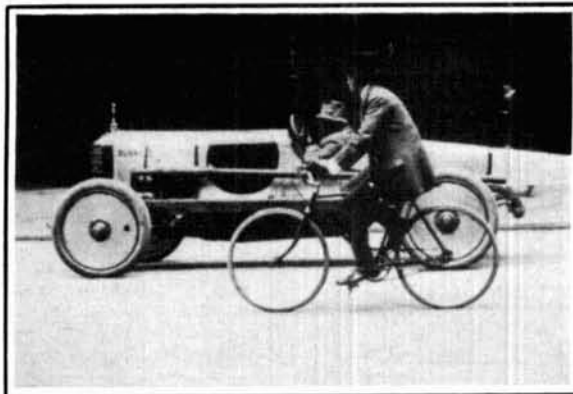
Dynamic range can then be usefully defined as two-thirds of the difference in level between noise floor and the intercept point, or alternatively as the difference between the fundamental response input level and the third order response input level as measured along the noise floor (see fig. 4). These are by no means the only methods of specifying dynamic range, and care must be taken in interpreting manufacturers figures. Using the above method of definition a dynamic range of 90 to 100 dB for 3 kHz bandwidth with an intercept point of 120 to 140 dB μV (+7 to +27 dBm), can be considered good. Note that because dynamic range — by definition — depends on the noise floor level, it will increase as bandwidth is reduced. Intercept point is sometimes specified (as are other parameters) in dBm, where a dBm is a dB relative to 1 mW into the system impedance (usually 50 ohms). 0 dBm (50 ohms) equals 224 mV PD (Potential Difference), or 113 dB μV , thus to convert from dB μV to dBm (50 ohms) simply subtract 113. (E.g., 0 dB μV = 1 μV EMF = 0.5 μV PD is -113 dBm). Note: dB μV is dB relative 1 μV EMF.

*Readers are urged to consider the outstanding intermod performance achieved by using FETs these days. Editor

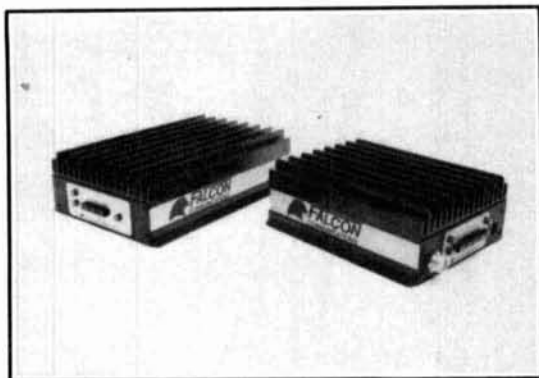
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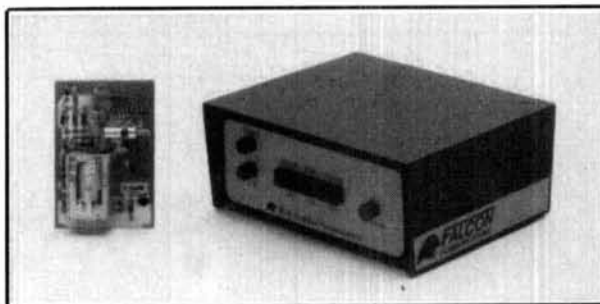
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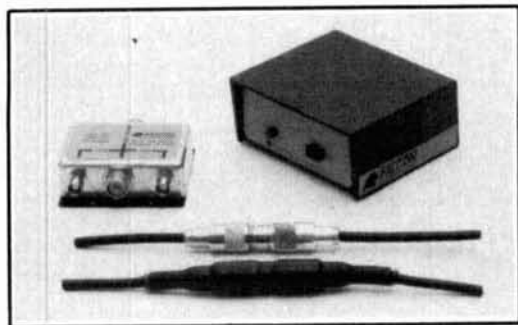
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2210G	220	130	10	265
2212	220	130	30	199
2212G	220	130	30	239
4410	440	100	10	225
4410G	440	100	10	265
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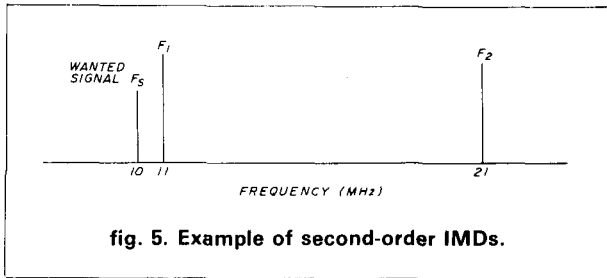


fig. 5. Example of second-order IMDs.

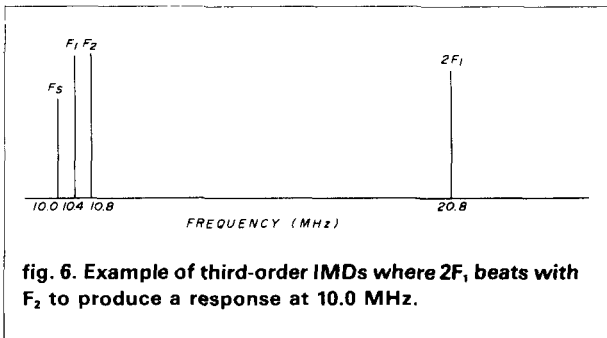


fig. 6. Example of third-order IMDs where $2F_1$ beats with F_2 to produce a response at 10.0 MHz.

intermodulation

Intermodulation product distortion (IMD) occurs when two large unwanted signals beat together (inter-modulate) in a non-linear receiver stage, to produce a product at the *wanted* frequency.

Second order intermodulation products equate to $F_1 \pm F_2$ (where F_1 and F_2 are the two unwanted frequencies). This process is shown in **fig. 5** where the two unwanted signals are at 11 and 21 MHz causing a beat at 10 MHz. (Signals at 6 and 16 MHz or 3 and 7 MHz would produce a similar product.)

One point to note about second order IMDs is that at least one of the unwanted signals must be outside the passband of any octave filter response that includes F_S and consequently can be rejected by any reasonably tight RF tuning, such as a good octave filter or sub-octave (less than an octave) filter.

Third-order intermodulation products are equal in frequency to $F_1 \pm 2F_2$. Thus both signals can be very close to the wanted signal, and well within the RF passband regardless of the type of RF tuning in use. This is seen in **fig. 6** where signals at 10.4 and 10.8 MHz produce a third-order IMD product of 10 MHz.

Intermodulation performance is typically specified as the levels of two unwanted signals not less than 20 kHz off tune (off the main signal frequency) to give a 0 dB μ V (1 μ V EMF) response. A good receiver will have a third-order IMD performance of 70 to 90 dB μ V. A statistical analysis based on data relating to actual signals received over the whole of the HF band on wideband (rhombic) antennas^{1,2} indi-

cates that at least 90 dB is required. As specified above, this corresponds to a dynamic range of 100 dB and an intercept point of 134 dB μ V or +21 dBm (50 ohms), as shown in **fig. 7**. It is easy to see why such a figure is needed. 90 dB μ V corresponds to 32 mV EMF and at almost any time there will be tens of broadcast (and other) stations putting between 10 and 100 mV onto a wideband antenna, with hundreds of others in the range 1 to 10 mV (EMF)!

Second-order IMD performance, often not stated, can sometimes be misleading. Second-order performance can be poor enough to present a real problem even with good third-order performance if a wideband antenna is used without an antenna tuning unit (ATU). This is due to the use of unshielded octal filters of less than perfect performance. Levels similar to third-order performance are *required*.

In-band intermodulation occurs when two signals *within* the IF passband beat (mix) to produce extra products. It is normally of little significance in HF communications except where multichannel "Voice Frequency Telegraphy (VFT)" systems such as "Piccolo" are in use. A typical level of performance for a good receiver is for a product of -40 dB with reference to two in-bands signals.

Cross modulation occurs when modulation from an unwanted signal transfers itself across and "modulates" the wanted signal. Again this is due to non-linearities in the early receiver stages, and sometimes the same modulation will reappear on each adjacent signal tuned in. Cross modulation is a third-order effect, so good third-order IMD performance will tend to mean good cross modulation performance (see **fig. 8**).

Cross modulation may be specified as the level required in dB μ V for a 30 percent modulated carrier greater than 20 kHz off channel to cause 3 percent cross modulation. A level of 70 to 90 dB μ V is considered good.

Blocking, or de-sensitizing, is similar to cross modulation, but in this case the large off-channel signal causes a reduction in wanted signal output. It is specified as the signal required to reduce wanted output by 1 dB. It can often be caused by a strong CW signal, causing gain to go up and down with the keying. 90 to 110 dB μ V is considered good performance, for a 1 mV (EMF) signal.

causes and cures of dynamic effects

Dynamic effects are caused by large off-channel signals driving the receiver into non-linearity. There are three fundamental methods of improving performance: (a) preventing the off-channel signals

a performance specification

This detailed specification indicates the performance required for a "very good" receiver. Additional parameters not mentioned previously, such as audio output power, have been added for completeness. Note that specifications like this should state worst-case figures (maximum or minimum values that are acceptable when the receiver is being tested); in some cases, however, "typical" values are used instead. This will at first sight make the receiver seem considerably better: for example, a worst-case NF of 10 dB will often (typically) be 7 dB. It is therefore important to establish whether a true specification or a list of typical values is meant. The specification below uses worst-case figures. Note dB μ V is dB relative to 1 μ V EMF.

specification

frequency coverage	50 kHz to 30 MHz (continuous coverage, using full synthesis)						
frequency display	7 digit, resolution 10 Hz						
reception modes	CW, FSK (RTTY), SSB (USB and LSB), ISB, AM, NBFM						
input impedance	50 ohms						
sensitivity	SSB, CW, ISB - 0.5 μ V EMF for 10 dB SINAD in 3 kHz						
(500 kHz to 30 MHz)	AM -	2.2 μ V EMF for 10 dB SINAD in 8 kHz, modulation 30 percent					
	FM -	0.8 μ V EMF for 10 dB quieting in 15 kHz bandwidth, 60 percent deviation					
		(Noise factor 10 dB)					
IF selectivity		response (kHz)					
	filter	available on	- 6 dB	- 60 dB	shape factor		
	15 kHz	FM	15.00	33.00	2.2		
	8 kHz	AM	7.80	14.00	1.8		
	2.7 kHz	SSB, ISB, CW	2.70	4.30	1.6		
	1 kHz	CW	1.00	2.00	2.0		
	300 Hz	CW	0.28	0.75	2.7		
dynamic range	104 dB. [Intercept point 140 dB μ V = +27 dBm (50 ohms)] (3 kHz bandwidth).						
third-order intermodulation	94 dB μ V. The levels of two unwanted signals both greater than 20 kHz off frequency will be 94 dB μ V [or -19 dBm (50 ohms)] to give a 1 μ V (0 dB μ V) response.						
second-order intermodulation	90 dB μ V. The levels of two unwanted signals both greater than 20 kHz off frequency will be 90 dB μ V [-23 dBm (50 ohms)] to give a 1 μ V (0 dB μ V) response.						
in-band intermodulation	-40 dB. Two signals in-band of equal amplitude will produce a product greater than 40 dB down.						
cross modulation	100 dB μ V. A 30 percent modulated carrier greater than 20 kHz off frequency must be 100 dB μ V [-13 dBm (50 ohms)] to cause 3 percent cross modulation.						
blocking	100 dB μ V. A signal greater than 20 kHz off frequency will be 110 dB μ V [-3 dBm (50 ohms)] to cause a 3 dB reduction of wanted 1 mV EMF (60 dB μ V) signal.						
reciprocal mixing	90 dB. An unwanted signal 50 kHz off frequency will be 90 dB above the level of a wanted on-frequency signal to reduce its SINAD by 3 dB, in a 3 kHz bandwidth.						
image rejection	90 dB μ V. The image frequency must be 90 dB μ V [-23 dBm (50 ohms)] to give a 1 μ V (0 dB μ V) response.						
IF rejection	90 dB μ V. For the first and second IF the level of an unwanted signal will be 90 dB μ V [-23 dBm (50 ohms)] to give a 1 μ V (0 dB μ V) response.						
crosstalk	-50 dB. On ISB mode the crosstalk between two equal 0 dBm (600 ohm) outputs shall be less than -50 dB relative to output at 1 kHz.						
response to internal spurs	All internally generated spurious signals are less than 3 dB above receiver noise. (3 kHz bandwidth.)						
stability	After 10 minute warmup, better than 1 part in 10 ⁸ /°C. Long term crystal aging less than 1 part in 10 ⁸ per day.						
antenna radiation protection	Less than 10 μ V _{pp} (20 μ V EMF) into 50 ohms (-87 dBm). (2pW) Receiver can withstand 30 volts at antenna input continuously. A spark gap is provided.						
AGC performance	For an input change of 90 dB, the output change will be less than 3 dB for all signal levels greater than 3 μ V _{EMF} (10 dB μ V).						
AGC time constants		SLOW			FAST		
		attack	hang	decay	attack	hang	decay
	ISB, SSB, CW	10 ms	2s	200 ms	5 ms	200 ms	50 ms
	AM, FM	20 ms	—	20 ms	5 ms	—	5 ms
BFO range (CW)	\pm 3 kHz						
AF outputs	Main output: 2W into 8 ohms at less than 3 percent Total Harmonic Distortion (THD) Headphones: 20 mW into 600 ohms Line: 600 ohms balanced line independent of AF gain control, settable to -10 to +10 dBm (600 ohms), i.e. 245 mV to 2.5 VPD.						
muting	0V on mute terminal for muting. Mute level at least 60 dB down.						
metering	Meter reads signal strength in dB μ V or AF line level in dBm (600 ohms).						
power requirements	200 to 250 VAC, 45 to 65 Hz, at 50 VA (Volt-Amperes)						
environment	Operating: 14 to 104°F (-10 to +40°C) Storage: -40 to 158°F (-40 to +70°C) Relative humidity up to 95 percent at 104°F (40°C) (non-condensing).						
MTBF (mean time between failure)	8000 hours						

from entering; (b) improving the linearity of the early stages of the receiver, prior to and including the roofing filter (a roofing filter is inserted in the first IF and is used to reduce the number of strong signals passing through the IF [chain]; it is not as narrow as the main selective filters); and (c) reducing the level of all signals.

This last method works because the response to the unwanted (dynamic) signals falls off at a faster rate than that of the wanted signals (see fig. 4). It is implemented by means of a front-end attenuator or by a wideband AGC loop (separate from the main AGC loop) which operates on the RF amplifier on large signals only, and thus can be thought of as being an automatic attenuator. Both methods have the disadvantage of reducing receiver sensitivity, and consequently other solutions should be found.

Method (a) involves the use of sub-octave filters or some sort of preselector tuning and can be very effective in reducing second order effects. However, as previously mentioned, third order products can be too close (in the signal passband) for tuned circuits to have an effect.

The only real solution is to improve linearity, (b). Bipolar transistors are particularly poor in this respect, but FETs are approximately square-law devices and are therefore very good in terms of third-order effects, but not as good for handling second-order products. Linearity can be improved by using higher power supply voltages and by keeping pre-roofing filter gain down to a minimum consistent with required sensitivity, and therefore by keeping

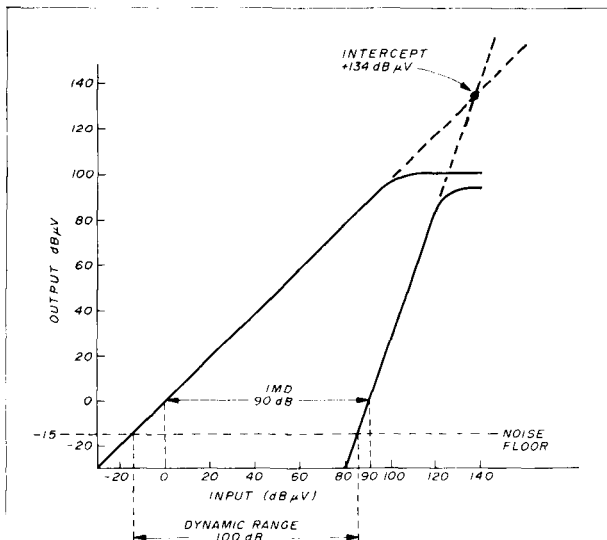


fig. 7. Good receiver third-order intermodulation distortion performance is shown as a 100 dB spurious-free dynamic range.

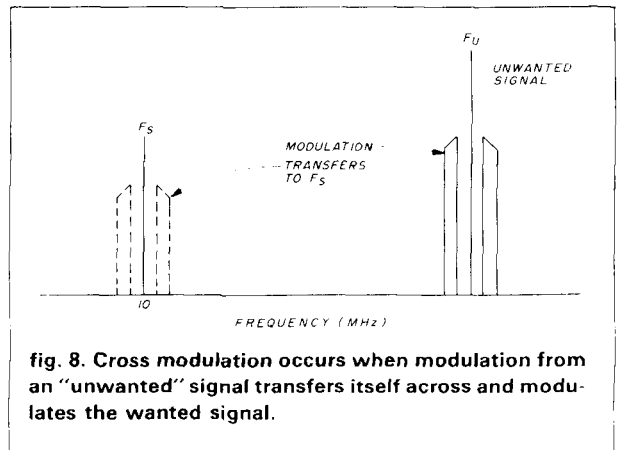


fig. 8. Cross modulation occurs when modulation from an "unwanted" signal transfers itself across and modulates the wanted signal.

noise levels down. In the extreme situation, the RF amplifier can be completely eliminated and the signals fed directly to a low-noise mixer via the front-end filters. This isn't such a drastic step as it might appear, because remember an NF of 15 dB is more than adequate on HF, and this can be achieved without an RF amplifier.

The mixer may be a double-balanced switching type diode mixer using high level LO injection to improve linearity. Components normally considered to be linear, passive, and reciprocal must be carefully checked to ensure that they are. This especially applies to ferrite cores used for RF coils and transformers; and crystal filters, which are often non-linear and non-reciprocal, i.e. having different characteristics if connected the "wrong" way around.

The practice of fitting protection diodes at the receiver input (often found on marine-band receivers) will also cause non-linearity, as will diodes used to switch filters. If all these points are carefully considered, very good linearity can be achieved with an intercept point of 140 dBµV or better. This level of performance ensures that IMD cross-modulation products are below atmospheric noise on HF.

Reciprocal mixing is due to high level unwanted signals mixing with the noise sidebands of the local oscillator/synthesizer, producing noise products at the receive frequency (see fig. 9).

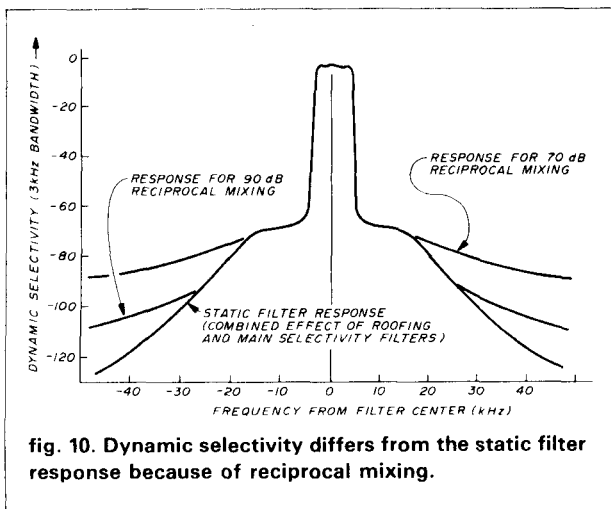
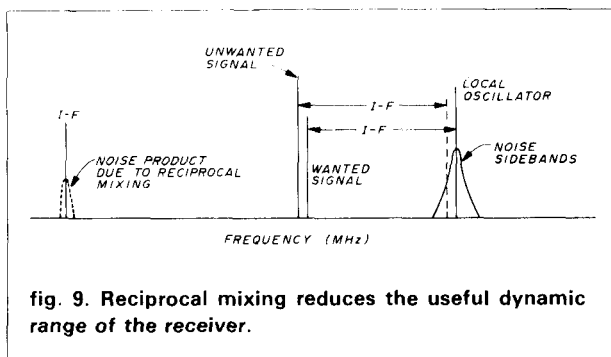
It is another phenomenon which until recently has been more or less ignored, partly because tube local oscillators are inherently "cleaner" than most of the modern solid-state synthesizers. It can be specified as the level in dB above a wanted signal that an unwanted signal can attain at a specified frequency off-channel (e.g. 50 kHz) at a specified bandwidth (usually 3 kHz) to reduce the S/N ratio of a wanted on-channel signal by 3 dB!

The oscillator noise can be reduced by employing

high "Q" (oscillator) circuits, and also by using high power oscillators, as the noise sidebands will then be relatively weaker, thus improving S/N. Phase locked loops can be very poor in this respect because they contain numerous noise sources which add together in the output, together with frequency jitter. In addition, PLL's frequently use low "Q" circuits and low power VCOs in the output. The noise produced is phase modulated and cannot be removed by limiting.

The action of reciprocal mixing in introducing off-channel signals into the IF at levels proportional to the distance away from the wanted signal (frequency separation) (see **fig. 9**) effectively reduces the selectivity of the receiver. This is shown in **fig. 10**, and the response curves indicate the dynamic selectivity of the receiver, that is, the selectivity of the receiver in a real signal environment. As can be seen, it's the stopband of the filter response that's been changed; with 70 dB reciprocal mixing, a considerable loss of performance occurs.

However, when the reciprocal mixing has been improved to 90 dB, its effect on filter response can be considered fairly minor. A frequency-synthesized receiver can achieve 90 dB, while a good (tube or FET) crystal oscillator can give 110 dB or better.



synthesizer noise and internal spurious responses

Internal spurious responses are responses of the receiver to signals generated within the receiver itself. These internally generated signals can be fixed (e.g., reference frequencies) or can move when the receiver tuning is changed. They cause problems when they occur at the signal frequency or an IF, and are generated by any oscillators within the receiver or by digital circuitry such as synthesizers and frequency counters. (A current trend is to use multiplexed fluorescent displays instead of DC driven LED displays, which generate low frequency interference. Adequate shielding must be employed to reduce radiation.) In using frequency synthesis techniques, many of the waveforms are digital, square waves with fast rise-times rich in harmonics. CMOS and LSI (N-MOS) is usually better in this respect than TTL which has faster rise times. Careful design, with adequate low-pass and bandpass filtering and with high "Q" output circuits, is important. With good design it is possible to keep spurious outputs 100 dB down on the main output level. This standard of performance should ensure that all spurious responses are no more than 3 dB above the receiver noise level.

stability

A fully synthesized receiver can have a stability equal to that of the frequency reference source.³ If an oven-temperature stabilized crystal oscillator is used, stability of less than 0.1 Hz/°C can be achieved. With partial synthesis the stability is governed by the stability of the VFO, but with cool, buffered solid-state designs it is possible to achieve long term drift rates of 100 Hz/hour with short term drift (even including lifting the receiver an inch and dropping it!) of 10 to 20 Hz. The latter is more than adequate for normal SSB/CW/RTTY/AM communications.

conclusions

The performance quality of the HF receiver has increased over the years and it is now possible to design a general coverage receiver that provides very high performance. As activity on the HF bands has constantly increased, this improvement in performance is of vital importance to maintain the ability to communicate.

Generally speaking, the cost of equipment has gone up in proportion to its complexity. There is, however, a reduction in real cost due, in part, to the availability of relatively low cost complex ICs, crystal filters, and FETs, as well as to improvements in design techniques.

Future trends will include more extensive use of

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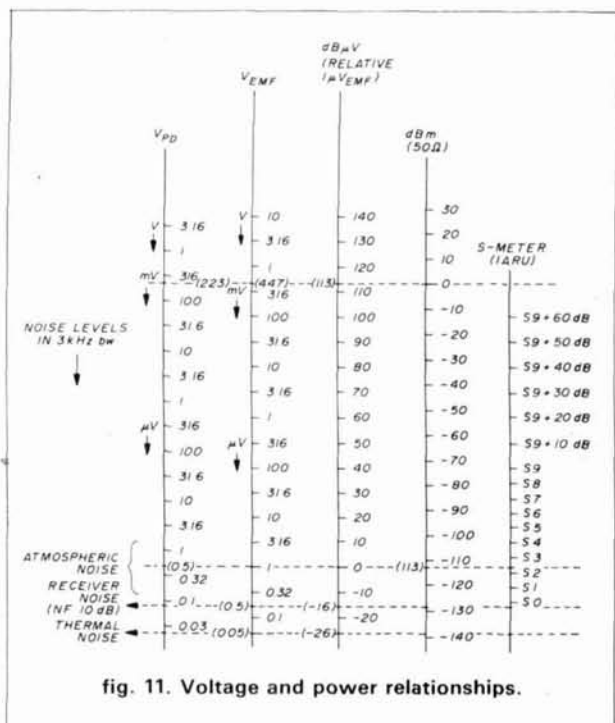


fig. 11. Voltage and power relationships.

ICs, with possibly a single chip full-synthesizer, and microprocessor control of the receiver. It seems likely that more extensive use will be made of remote control of receivers via data link, and that in communications centers a central mini-computer will be linked to each operating position, performing a variety of useful functions. On the domestic scene, the home computer can be linked to the receiver via an RS232 line, and can be used to decode RTTY and SSTV signals, etc. It could also be used as a big "memory" to store channels (frequencies/modes/filters and a channel ident) for instant recall.

In conclusion, receivers available now offer performance that ten years ago could be obtained only from professional receivers costing ten times as much, and which twenty years ago could not be obtained at all.

appendix

Fig. 11 shows the relationships between levels as specified in V(EMF), V(PD), dBμV, and dBm (50 ohms). Also shown is the S-meter response as recommended by the IARU, the International Amateur Radio Union. This specifies that S9 should be at 50 μV PD, or 100 μV EMF; and that each S-point should be at 6 dB intervals. Also shown are the thermal noise threshold and typical receiver and atmospheric noise terms for a 3 kHz bandwidth.

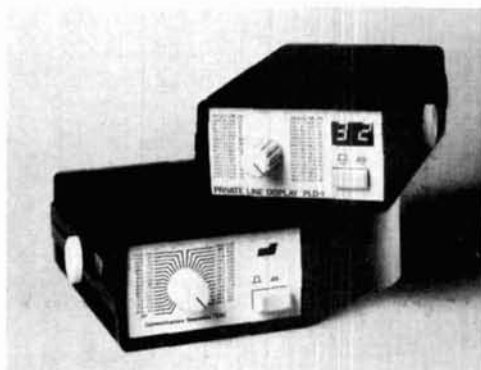
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3. J.A. Dyer, "High Frequency Receiver Design," *Radio & Electronics World*, February, 1983, pages 48-53.

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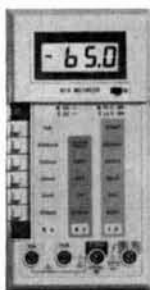


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VHF/UHF antennas and antenna systems

There is probably no other VHF/UHF topic that inspires such a vigorous line of conversation than the subject of antennas. Hardly a day goes by when I'm not asked questions like "What type of antenna do you recommend?" or "What's the difference between this antenna and that one?" The reason we're so interested in this subject is that there is no better way to improve the performance of your station: *a 1 dB improvement in antenna gain yields a 2 dB overall increase in your station capability, since we gain 1 dB on transmit and 1 dB on receive.*

The answers to the questions above are neither simple nor obvious; they take time to answer thoroughly. Furthermore, the answers may not always involve performance as much as economic or structural considerations. With this in mind, I've decided to discuss the popular types of VHF/UHF antennas and the tradeoffs between them in this month's column. Future columns will zero in on ways to measure and obtain peak performance with the antennas we are using.

antenna types

The three major types of VHF/UHF antennas presently in use (excluding FM/repeater types) are the collinear array, the Yagi and the parabolic dish. Each type has its own advantages and disadvantages. While all can produce high gain, they differ

vastly in form factor. Let's discuss each type individually and see what each can and can't do.

collinear array

The collinear array was very popular among VHF/UHF'ers and especially EME'ers (Earth-Moon-Earth) on 2 meters and 70 cm (432 MHz) before high performance VHF/UHF Yagi an-

tennas were designed. It usually consists of a group of half-wavelength dipoles in front of a screen or set of reflectors (fig. 1). Technically speaking, you could probably call this an array of 2-element Yagi antennas. Some individuals have even placed additional directors in front of the driven dipoles in an attempt to increase gain.¹ The unique thing about a collinear array is that the feed system is usually some form of open wire line. The collinear, unlike many antennas, is usually quite broadband. Efficiency can be very high and gain is mainly a function of the size of the array. The extended-collinear is a stretched-out version that has fewer elements and higher gain approaching 80 to 95 percent efficiency.² This antenna is treated in depth in reference 2 and improvements are mentioned in reference 3.

Collinear antennas offer several advantages: They are not "critical" to build, are usually low in cost, have a low loss feed system, and when used on EME, can be readily adapted to polarity rotation. They are also easy to array when extremely high gain is desired. The principal drawback is their overall size, which usually precludes mounting other antennas on the same mast. (I'll discuss this subject later in this column.) There may also be problems in areas where moisture is present, since the VSWR may increase if the feed system gets damp or wet.

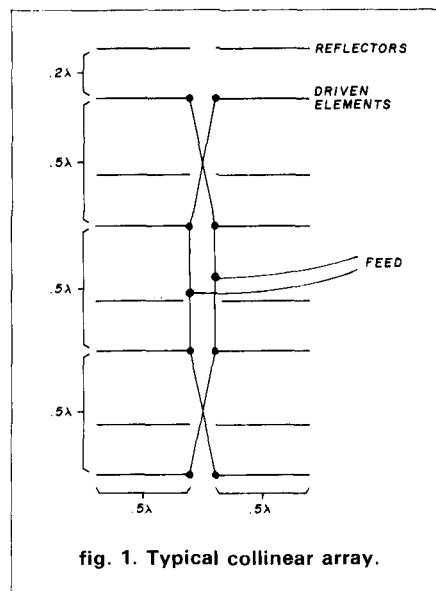


fig. 1. Typical collinear array.

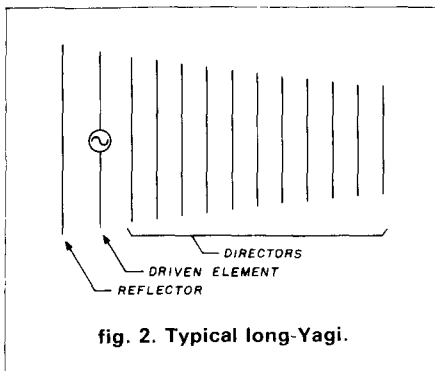
tennas were designed. It usually consists of a group of half-wavelength dipoles in front of a screen or set of reflectors (fig. 1). Technically speaking, you could probably call this an array of 2-element Yagi antennas. Some individuals have even placed additional directors in front of the driven dipoles in an attempt to increase gain.¹ The unique thing about a collinear array is

Yagi and Yagi type antennas

The Yagi antenna (fig. 2) is particularly popular where space is at a premium or when only narrowband operation is required. It is presently the workhorse in the VHF and lower UHF spectrum. The first high gain VHF Yagi designs were published by Kmosko and Johnson,⁴ Greenblum,⁵ and Ehrenspeck and Poehler.⁶ Unfortunately, these Yagis weren't always

as good as claimed, were hard to duplicate, and when duplicated, often failed to deliver the promised gain. Some had very high side lobes.

The WØEYE 70 cm 4.2 wavelength 15 element Yagi⁷ with corrections⁸ was first published in 1972. Based on the unpublished work of Pete Viezbicke at NBS (National Bureau of Standards), it was the first really high gain Yagi with a clean pattern that was easily duplicated. Don Hilliard, WØEYE (now WØPW), and I urged Pete to publish his work and he finally did so in December, 1977, in *NBS Technical Note 688*, now out of print.⁹ This publication, the result of extensive studies done by NBS in the 1950's to develop high gain arrays for ionospheric scatter, included six different models with boomlengths of 0.4 to 4.2 wavelengths. In the August, 1977 issue of *ham radio*, I published an article which included the



majority of the NBS material along with several practical examples of Yagi designs using this material.¹⁰ (A future column will deal with this topic in greater detail.)

The NBS Yagi designs described above work well through 70 cm, but the longest one is only 4.2 wavelengths (approximately 10 feet or 3 meters at 70 cm). Günter Hoch, DL6WU, has been working on this problem and has recently produced some very long Yagis¹¹ based on an extension of the work of Greenblum.⁵ Hoch has been able to increase the gain approximately 2.35 dB each time the boomlength is doubled and has demonstrated this

success even at 23 cm (1296 MHz). I have verified his information with a 70 cm 9.25 wavelength (21 feet or 6.5 meters) model which measured greater than 17.5 dB over a dipole gain at the 1981 Central States VHF Conference.¹² At the 1983 Central States VHF Conference, KL7WE entered an even longer (24 feet or 7.5 meter) 70 cm Yagi model based on the same information and it measured approximately 18 dB over a dipole. Hence the quest for extremely high gain Yagis is finally showing promise.

the Quagi

In April 1977 Wayne Overbeck, K6YNB/N6NB, published a description of an antenna he called the "Quagi."¹³ This antenna is basically a Yagi with a cubical quad-type reflector and driven element. The design shown was low in cost, used a wooden boom, and was fed directly with coax. Newer designs have been published;¹⁴ DL9KR and others have further optimized this design and modified the feed systems for use in large (groups of 16) EME arrays. This design could probably still use some optimization; only a few specific designs are available.

the log-periodic antenna array

The log-periodic antenna array,¹⁵ a series of elements resembling a Yagi antenna, but with all elements fed by a special feed system, has never gained much acceptance among VHF/UHF'ers (except ATV'ers) because it is basically a wideband (multi-octave) structure with only moderate gain. The late Oliver Swan developed a hybrid antenna called the log-periodic Yagi, which has somewhat wider bandwidth than the typical Yagi.¹⁶ It uses a log-periodic feed system to excite the directors of a Yagi structure. A 70 cm log-periodic Yagi has also been published.¹⁷

loop Yagis

Until recently, very few Yagi designs were used on the 23 cm band because they were hard to duplicate

and extremely tight tolerances were required. The boom-to-element attachment had been a problem both electrically and mechanically, which further aggravated the tolerance problem.

In 1974, Mike Walters, G3JVL, decided to take a different approach to the Yagi antenna. At first he experimented with cubical quad loops made of wire, but could not obtain the desired performance. He then changed the elements to thin 0.028 inch (0.7 mm) aluminum straps 3/16 inch (4.7 mm) wide bent in a circle. After much trial and error, he developed a 23 cm, 28 element high gain "loop Yagi."¹⁸ Later he developed longer models and also a table of corrections for different boom and element sizes.¹⁹ The use of a round boom with his unique element-to-boom mounting method and the correction tables has been one of the reasons this design has been so successful. This antenna has been used as low as 70 cm and as high as 3 cm (10.256 GHz). A 45-element W1JR-designed loop Yagi had greater than 19 dB gain over a dipole as measured at the 1983 Eastern VHF/UHF Conference.

stacking

When really high gain is desired (such as in EME), Yagi antennas can be either arrayed or stacked. Generally speaking, every time you double the number of Yagi antennas in an array, you can increase the gain approximately 2.5 dB if the proper stacking distance is used and the feedline loss is kept low (this will be covered in a later column). Recent work on 70 cm EME tends to confirm that the best compromise is to use the highest gain Yagi design possible. Then the number of Yagis in the array will be at a minimum, and feedline problems will be more manageable. For general terrestrial operation, it is usually desirable to stack antennas vertically so that the beamwidth in the azimuth plane will stay wide and thereby spare you severe rotator aiming problems. This is particularly desirable for

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meteor scatter communication, in which the field of reflection can be wide.

parabolic dish antennas

The parabolic dish (fig. 3) is still one of the favorite antennas on 23 cm and above — especially on EME. The gain of this type of antenna is primarily determined by its size (diameter), providing that the feed or illuminator is properly chosen. There are many advantages to using this type of antenna. It is frequency-independent, meaning that you can change frequency by simply replacing the feed

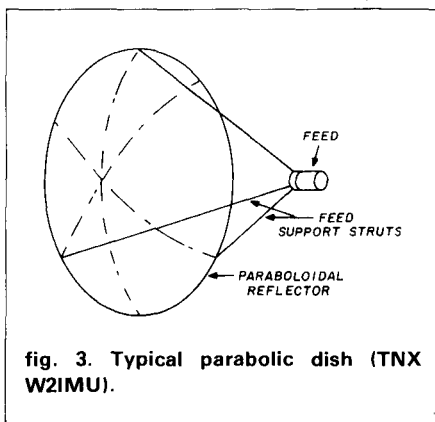


fig. 3. Typical parabolic dish (TNX W2IMU).

system with one for the proper frequency. A properly built parabolic dish antenna system will also be quieter on receiving because side and rear lobes are lower than most other antenna systems. Its principal disadvantages are that efficiency is low (typically 50 to 55 percent, even if everything is done right) and high wind resistance. (Recommended feed systems for parabolic dishes are described further in reference 3.)

conclusion

Hopefully the above information and the additional articles referenced will make it easier for you to determine which antenna type is best for you. The collinears are usually low in cost and easy to build, but if high gain is desired, they must be large. Where multiple band capability is desired on a single mast, the Yagi is hard to beat. When high gain on UHF and especially EME are sought, the

parabolic dish offers many advantages; but again, if high gain is desired, the antenna system must be large. While we've discussed many types of antennas for VHF/UHF, we haven't discussed them in great detail. We'll deal with each type more comprehensively as time passes. But it will be up to you to weigh the advantages and disadvantages of each, make your decision, and take the plunge.

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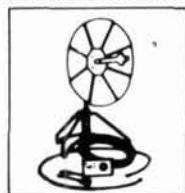
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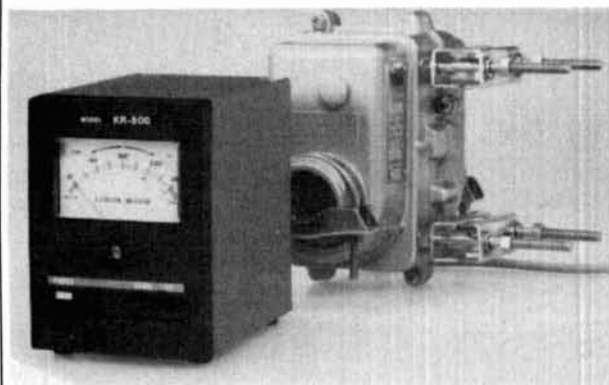
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transmitter tuning aid: buffer your load with this resistive network

The temporary high SWR often encountered when tuning up your transmitter and tuning unit can often lead to disastrous consequences: abnormally high voltages and/or currents can easily damage expensive equipment, even in only a few seconds. If you once, perhaps, pit the variable capacitors of your final amplifier by an arc over, there is very little that can easily be done to repair the damage. In some cases, manufacturers have designed automatic circuitry that reduces power output when the SWR goes up, but this technique offers only limited protection rather than a basic solution.

With these concerns in mind, I developed and built a dummy load to overcome these problems. All discussions and calculations are based upon a 50-ohm impedance system, which is common to most of our modern transmitters and tuning systems. However, the equations can be used for any impedance system, or for any SWR limiting value desired.

The basic idea is to use a balanced-T resistive network, with a characteristic impedance of 50 ohms, to absorb most of the transmitter power during tune up, while still providing sufficient power to allow for proper tune up of your transmitter and antenna system. The isolation effect of the network is such that it limits your transmitter SWR to a maximum of 2:1, even if you were to short out or open the input to your tuning unit, which would be the worst condition that could develop.

Fig. 1 shows the basic circuit, along with theoretical calculated values of resistance, and wattage dissipation for a transmitter with an output of 100 watts. This figure was chosen for ease of calculation and presentation of the developed equations. Although the network is a balanced type, so that R_A is numeri-

By William Vissers, K4KI, 1245 South Orlando Avenue, Cocoa Beach, Florida 32931

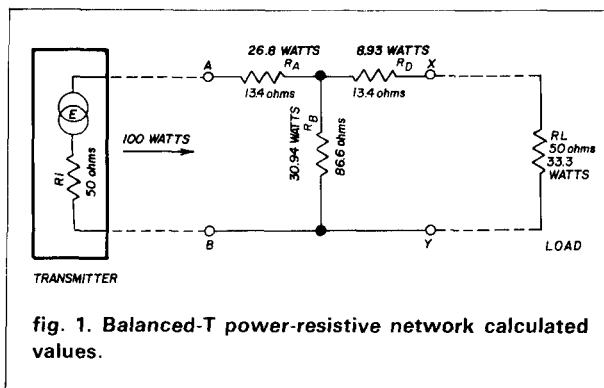


fig. 1. Balanced-T power-resistive network calculated values.

cally equal to R_D in ohms, the different subscripts are shown to denote different wattage dissipation.

The equations relating the maximum value of SWR, which was chosen as 2:1, the characteristic impedance of the system R_C of 50 ohms, and the network resistance values are:

$$R_A + R_B = (R_C)(SWR) = (50)(2) = 100 \text{ ohms}$$

$$R_A + (R_D)(R_B)/(R_D + R_B) = R_C/SWR$$

$$= 50/2 = 25 \text{ ohms}$$

Having two equations and two unknowns, the network is solved using the basic quadratic equation, where:

$$R_A = R_D = 13.4 \text{ ohms and } R_B = 86.6 \text{ ohms}$$

Again looking at fig. 1, if terminals X-Y are shorted out, the resistance looking in at points A-B is 25 ohms, so the transmitter SWR would be $50/25 = 2:1$. If terminals X-Y are left open, the resistance looking into points A-B would be 100 ohms, and the transmitter SWR will once again be $100/50 = 2:1$.

The characteristic impedance of the network is determined by these open and short circuit conditions where:

$$R_C = \sqrt{(R_{s.c.})(R_{o.c.})} = \sqrt{(100)(25)} = 50 \text{ ohms}$$

All of the original design requirements have been satisfied.

During actual tune up, R_L , the 50-ohm load resistance, is replaced by the input of the matching unit in the circuit in fig. 2.

From fig. 1, it is seen that the power delivered to the load is 33-1/3 watts, or 1/3 of the transmitter output power of 100 watts at terminals A and B. This still provides plenty of power for tune up, yet offers an added advantage in that your "tune up QRM" has dropped by a factor of $(10)(\log_{10})(1/3) = -4.77 \text{ dB}$. This reduction will be appreciated by other Amateurs on our already crowded bands.

A few years ago I experimented with building and using ordinary two-watt carbon resistors in parallel,

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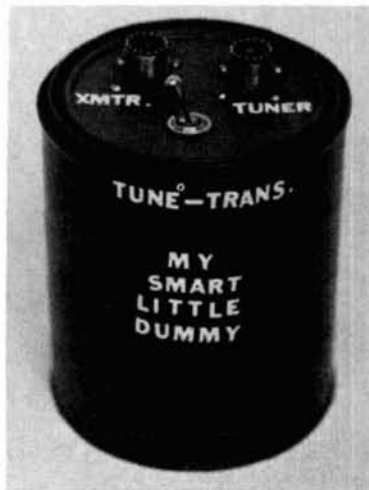
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immersed in oil for a simple 50-ohm dummy load. I decided to follow the same procedure used then for the final T network, (fig. 2), in this project. My earlier tests had shown it was possible to use a heat dissipation factor of 2.5 as a conservative approach for normal tune-up time. To choose resistors, I checked to see what was available at my local electronics store, and then calculated what was needed to approach the theoretical values previously identified. For R_A , six 82-ohm, 2-watt carbon composition resistors in parallel equal 13.67 ohms with an allowable wattage rating of $(\text{watts per resistor})(\text{number of resistors})(\text{derating factor}) = (2)(6)(2.5) = 30 \text{ watts}$, which was above the previously calculated value of 26.8 watts. (The values of resistors specified are shown in the parts list of fig. 2.)

Although the actual network resistance values differ slightly from the theoretically calculated values, actual tests showed the differences to be insignificant in operation. The network is very tolerant of small resistance deviations. This means you can simply use off-the-shelf items — there's no need to handpick resistance values.

There are other advantages in using the unit. Because your transmitter SWR during tune up can never go above 2:1 under the worst of conditions (and even then your reflected power is still only 11 percent of forward power), your final amplifier output tuning control positions will be very close to their actual positions when you are finally tuned up to a perfect SWR of 1:1. This makes tune up faster than it would be if you tried to tune up your amplifier while it was connected directly to your matching unit. Under



Completed balanced-T resistive network.

the latter circumstances, the operation of tuner and transmitter controls is often so complex as to make one wish for four hands instead of two, and an extra pair of eyes to watch the SWR meter, lest it go much over 2:1.

Another advantage of the unit is that you can switch it from TUNE to TRANSMIT when you are tuned up without having to turn off transmitter power, the circuit design prevents dangerous switching transients from occurring. Finally, because the unit is an in-line device, you will not need an expensive coax switch as might be necessary with a regular dummy load.

Although the design shown was for a balanced-T network, an equivalent Pi network with suitably calculated resistances and wattage values works just as well. The technique is also adaptable for high power. Actually, my first unit built was a converted home-brewed, flea market, 1 kW gallon-sized dummy which used a number of vitreous enameled non-inductive resistances, similar to the ones manufactured by Dale Electronics, Inc., P.O. Box 609, Columbus, Nebraska 68601. Although my copy of their catalog did not specify the highest frequency at which their non-inductive resistors could be used, my old re-structured dummy, with similar resistances, worked fine down through 10 meters.

This device has also been thoroughly tested by Russ Forsyth, K4YS, under a variety of operating conditions. His helpful suggestions and comments are part of this article.

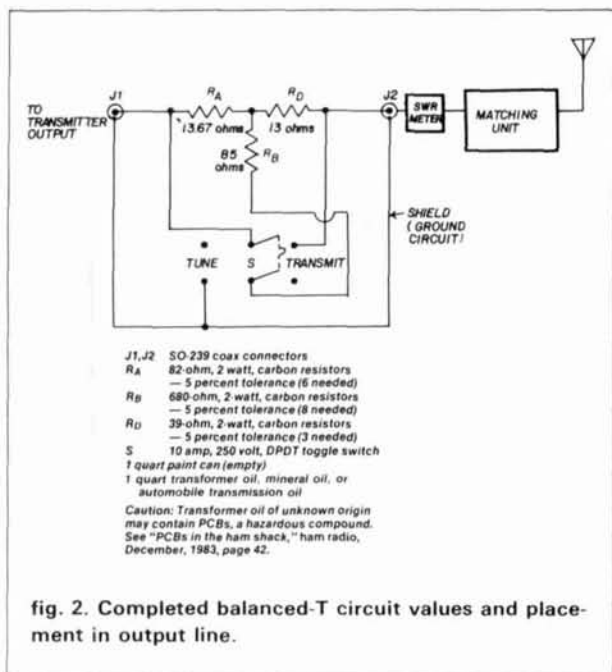


fig. 2. Completed balanced-T circuit values and placement in output line.

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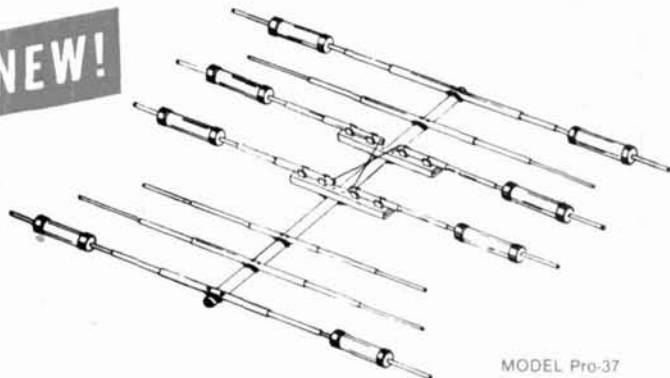
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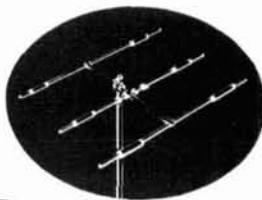


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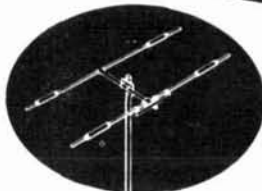
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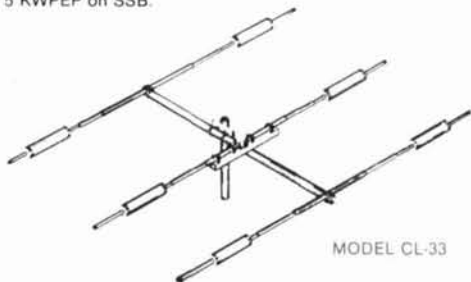
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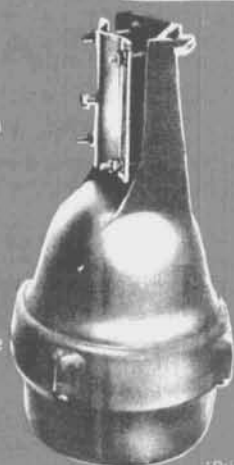
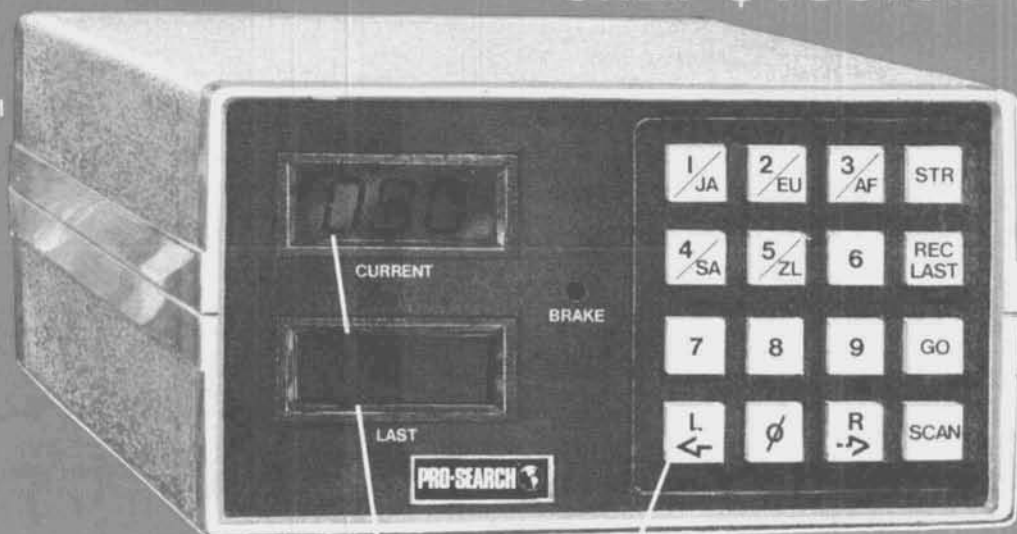
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better-sounding SSB

If you want your SSB transmit audio to sound like your voice, this article was written for you.

I have never accepted the notion that good audio died with AM. After all, isn't SSB just AM with two of the unnecessary parts removed? The question is, why doesn't SSB audio give you good audio quality all the time? Armed with some electronic test equipment and a yearning for nice sounding audio, I set out to find the answers.

The results of my research indicate that three criteria must be met to achieve transmit audio quality that compares favorably with AM: the first is smooth frequency response; the second, a wide enough passband or "window" to be able to include most of the important frequencies in a human voice; and third, distortion should be low enough so that the voice does not sound rough.

generating SSB

In most SSB transceivers the mixing of the voice (audio) frequencies and the RF takes place in a balanced modulator. The balanced modulator is really a mixer that combines the audio and RF together. However, mixing is not an accurate way to describe the process, since the output is the sum of the audio frequency and the RF, or carrier frequency, and the difference of the audio frequency and the carrier frequency. (The name "mixer" should probably have been "adder-subtractor.") The *sum* output is upper-sideband (USB). The *difference* output is lower-sideband (LSB). Since both USB and LSB come out together on the sole output terminal, the result is

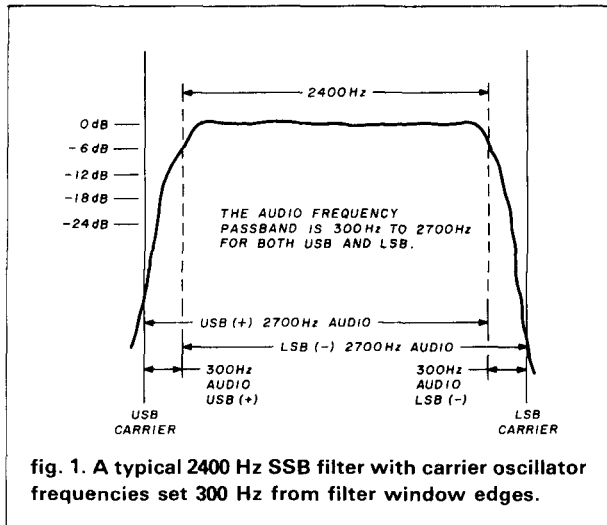
double-sideband (DSB). Because of the balanced design of the modulator, neither the audio nor the carrier oscillator (input) signals should appear in the output.

It is important to remember that USB is always an addition process, and that LSB is always a subtraction process. For example, with a carrier frequency of 10,000,000 Hz and an audio frequency of 440 Hz, the USB output is 10,000,440 Hz and the LSB output is the difference or 9,999,560 Hz. If a second audio frequency were also fed into the balanced modulator along with the 440 Hz signal, then two frequencies would appear on USB and two frequencies would appear on LSB. If 1000 Hz were included with the 440 Hz signal, the USB would contain two frequencies: 10,000,440 Hz and 10,001,000 Hz. LSB would also contain two frequencies: 9,999,560 Hz and 9,999,000 Hz. When a voice is applied to the input of the balanced modulator, the output becomes more complicated, since a human voice contains many frequencies, but the additive and subtractive processes still apply.

producing just one sideband

The most popular system of selecting only one sideband incorporates a filter to reject the unwanted sideband and pass the desired sideband. It is common practice to use only one filter for both sidebands. In order to do this, two different carrier oscil-

By Richard L. Measures, AG6K, 6455 La Cumbre Road, Somis, California 93066



lator frequencies are used. On USB the carrier oscillator is placed about 300 Hz lower than the low frequency edge of the filter window. Since USB adds the voice and carrier frequencies, the sum will be higher than the carrier frequency and any audio frequency of 300 Hz or more will create USBs that can pass through the filter window. On lower sideband, the carrier frequency is placed about 300 Hz above the upper edge of the filter. Since LSB subtracts the audio frequency from the carrier frequency, an audio frequency of 300 Hz would just reach the upper edge of the window. In either case an audio frequency of 200 Hz would be mostly rejected since it could not add or subtract enough to each carrier frequency to reach the window. If the carrier oscillators were reset to a point 200 Hz from the edges of the window, then 200 Hz would produce a signal that will pass through the window. It is important to remember that the audio frequency itself does not pass through the window. The signals that pass through the filter are at a radio frequency, usually around 9 MHz. These signals are offset from the carrier frequency by the audio frequency. In order to receive SSB, the signal is combined with the same carrier frequency in the receiver as was used originally in the transmitter. The combining is done in another mixer called a product detector which produces audio output. This is the reverse of the process that began in the balanced modulator.

what voice frequencies are needed?

In the above example with the carrier oscillator frequencies set 300 Hz from the upper and lower frequency edges of the filter, it was shown how a 300 Hz audio signal would just reach the filter window. The next point to consider is how high in frequency

can the audio signal go without missing the filter window. This is an important consideration because human speech contains high frequency sounds that are made by using the tongue to direct airflow against the teeth. This group of sounds, known as sibilants, includes S, soft C, F, and H. To do real justice to human speech, the window should allow at least 2800 Hz to pass. Allowing only 2500 Hz to pass makes a noticeable difference in the clarity of the sibilants. If a standard 2400-Hz wide filter is used, then the opposite side of the window is 2400 Hz away from the edge of the filter where the frequency generated by the 300 Hz audio signal would pass (fig. 1). The highest frequency audio that could pass without falling out of the window would be 300 plus 2400 or 2700 Hz. This audio would be acceptable, but would have more sibilance if the carrier frequency were moved 100 Hz farther from the filter window edge. This would move the audio passband from 400 Hz to 2800 Hz. This audio would be pleasing to most people with normal hearing. If a wider filter window were used — 2700 Hz — the audio quality would sound even better, since the carrier oscillator frequency could be reset so that the full range of important human speech frequencies — 300 Hz to 3000 Hz — could pass through the 2700 Hz window.

linearity

If the audio sounds rough and splatters on other frequencies, then linear operation is not occurring. The human ear can detect low levels of distortion. Some musicians can hear less than one part distortion in 10,000. That is -40 dB. I don't notice on-the-air distortion if it is -38 dB or better. If a station 4 kHz down the band has a 30 dB over S9 signal and his distortion products are down only 30 dB, you are going to see close to S9 splatter on the window you are listening to.

Most of the radios on the market today are fairly clean. The cleanest by far are those with vacuum tube finals employing RF negative feedback. If operated conservatively, some of the solid-state radios can also deliver a clean sound.

On my solid-state transceiver I was able to achieve -40 dB distortion products on my voice, but only after carefully setting the driver and final transistor idling current to the values called for in the service manual. I could only maintain the -40 dB distortion level if I kept the ALC level in the bottom quarter of the ALC scale. If I tried to operate with the ALC at the top of the "safe" range, the distortion increased by a factor of nine.

This test does not apply to the common two-tone method of measuring distortion. A human voice is much more complex than two tones and provides a

more accurate and severe performance test. Blowing into three holes of a harmonica also works well.

The mechanism of non-linearity is essential if a mixer is to do its job of adding and subtracting. In an amplifier, non-linearity means that the amplifier is going to do some unwanted adding and subtracting of any two or more frequencies fed into the amplifier input. Human voices have many frequencies. If the linear amplifier is not linear, there will be more frequencies, and some of them are likely to land on someone else's conversation on a nearby frequency.

measuring distortion

Distortion can be measured with a separate receiver, provided it has an S-meter of known accuracy. The linearity of the S-meter is the essential factor in making this test. The number of microvolts that equals S9 does not matter. Use a short clip-lead for the receiver antenna and a 20 dB attenuator. Tune to the frequency on which you are transmitting and watch the S-meter while you speak into the microphone. Note the S-meter reading and retune the receiver either 4 kHz above or 4 kHz below the transmitter frequency. Speak again and note the S-meter reading. The dB difference is your distortion level. A reading of -40 dB is excellent. A reading of -30 dB is not quite anti-social, but -20 dB is definitely in the "skunk" category. The all-time record worst distortion I have witnessed was -9 dB — this is *not* the way to win friends. Measuring distortion as described works best on your own transmitter. It is possible to use the same method on the air to measure someone else's distortion, but fading signals can make measurements inaccurate. To do this, a real antenna is more useful than the clip lead. I recommend 4 kHz spacing to ensure that your receive window does not appreciably overlap the other operator's transmit window. If the windows touch, the measurement will mean nothing, since you won't be able to separate the distortion from the desirable audio frequencies.

design defects

During my first SSB experiments I was amazed to learn that listeners could detect a small change in flatness of frequency response. If your transmitter has a 3 dB loss at the high end, the lack of sibilance is noticeable. Many of the modern rigs on the market today have a built-in rolloff of the high frequency transmit and receive audio. The amount of high frequency (treble) rolloff varies from 4 dB to 10 dB in the rigs that I have measured. Usually the receivers have the more uneven audio response.

The usual culprit in the transmitter section can be found at the collector of the microphone amplifier. In my transceiver there was a 0.015 microfarad capaci-

tor from the collector to ground. About 60 percent of my 2900 Hz audio was going to ground at this point. I replaced this with a 0.001 microfarad capacitor and the treble passed through unscathed. Since the purpose of this capacitor is to prevent your own RF from getting into your audio, a non-inductive capacitor such as a disc ceramic type is advised.

I noticed that the high frequency audio in my receiver seemed attenuated. My test equipment confirmed a 9.5 dB rolloff at 2900 Hz. This meant that 11.2 percent of the audio was making it to the speaker at that frequency and that 88.8 percent was not. The culprit in this case was a large-value capacitor at the output of the product detector. The value used was 0.033 microfarad. This capacitor had a reactance of only 1600 ohms at 2900 Hz. The audio impedance of the product detector was about 5000 ohms. Shunting a 1660 ohm capacitor to ground is bound to be a spoiler. The purpose of this capacitor is to let the audio pass and to suppress the RF energy. Again the capacitor is necessary — but the value must be picked with some prudence. (Bigger isn't always better.) I changed the capacitor to 0.0033 microfarad. The audio sparkled. Voices were easier to understand.

microphones

Anyone who reads the advertising for microphones knows that you need a special microphone for SSB. It sounded reasonable to me, so I bought two microphones with lots of "punch." One was a crystal element type. The other microphone had a rising response dynamic element. Both microphones had more highs than lows — just the ticket to "get more DX." I also owned a flat response electret condenser microphone.

Early one winter morning I was working a DX station on 80 meters. I asked the DX station if he would listen to three different microphones and select the one that was easiest to copy under adverse conditions. I tried microphone number 1, number 2, and number 3. The microphone that produced the best copy was microphone number 3 — the condenser microphone — and the same microphone my local friends preferred to listen to. To be sure, this was only one test, but it certainly cast doubt on the traditional view of what it takes to be understood on SSB. Perhaps we don't really need to sound like ducks after all. Many amateur radio transceiver manufacturers now offer condenser microphones. Most of the microphones worn by television newscasters are of this type.

adjusting the carrier oscillators

The most important adjustment in an SSB transceiver is the carrier oscillator. This adjustment sets

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the position of the audio frequency window on both RECEIVE and TRANSMIT. The only test equipment needed is a common audio generator that can be adjusted for a few millivolts and an RF wattmeter-dummy load. Connect the generator to the microphone input. Adjust the audio frequency for 1000 Hz. Set the level at a value that is about half of the maximum output power — for example, 40 watts. Adjust the generator higher in frequency until the wattmeter reads one quarter of the wattage produced when the audio frequency was 1000 Hz. This 25 percent power point is the minus 6 dB rolloff. Move the audio frequency lower until the power drops to 25 percent again. All of the frequencies between these two audio frequencies comprise your transmit window. If you find that your high frequency term is less than 2800 Hz at the rolloff point, your carrier needs to be moved farther away in frequency from the filter window. The low frequency rolloff should be 300 Hz or higher, but not above 400 Hz or it will sound thin (or tinny). If you are using a 2000-Hz wide filter you have a problem unless you want to sound like a 1910 model telephone. Wider filters are available for many rigs, direct from the manufacturer. I bought 2700 Hz filters for both of my transceivers.

Some people believe that a wider filter will impair reception because of increased splatter from adjacent stations. Most of the splatter on the ham bands is due to distortion products coming from the other fellow's transmitter; distortion products such as these can never be eliminated with a receive filter because the distortion is on the same frequency as the station you are trying to listen to.

There are devices on the market that will correct for a slight loss in treble transmit response caused by a 15-cent capacitor being the wrong value. But if the carrier oscillator is misadjusted, these devices will not be effective. For example, if the carrier oscillator is set so that the high frequency rolloff is 2200 Hz, the SSB filter is going to have the last word on what gets through. Boosting the treble with an accessory can make the problem less noticeable, but will not restore the high frequency rolloff to 2800 Hz as it should be.

Carrier oscillators can also be adjusted by ear. The rule is to simply set it to where it sounds good. While this may take a little longer than it would using test equipment, the method works. The only problem lies in finding a listener who is not afraid to tell you what you really sound like so you can know which way to move the adjustment. Some people don't want to hurt your feelings; they don't like to give a "bad" report.

Clean audio is a pleasure to listen to on the air, and rotten, splattering audio is a blight on the bands.

ham radio

ham radio TECHNIQUES

Bill W6SAI

the groundplane loop antenna

Every once in a while an interesting antenna comes along that causes me to ask, "Why didn't I think of that idea?" A case in point is the groundplane loop antenna described by Hans Wuertz, DL2FA, in *cq-DL*, the monthly publication of the German Amateur Radio Club. An English translation of the pertinent data on the loop was recently published in *Radio Communication*, the journal of the Radio Society of Great Britain (RSGB).

The basic configuration of the DL2FA loop is shown in **fig. 1**. The design is a half of a full-wave loop antenna with the other half formed by the ground image. With a 0.2 wavelength half-loop for the lowest operating frequency, DL2FA claims that the antenna can be tuned to cover a 2-to-1 frequency range (3.5 to 7 MHz, for example).

Even though the radiation resistance of the antenna is low, the efficiency of the antenna is excellent, and at 3.5 MHz the field strength is only 2 dB less than a full-size quarter-wave vertical antenna. At 7 MHz, the field strength is within 0.5 dB of a full-size vertical.

To achieve low loss, the loop and ground return are constructed of 3/8-inch (approximately 10 mm) diameter copper tubing. The copper

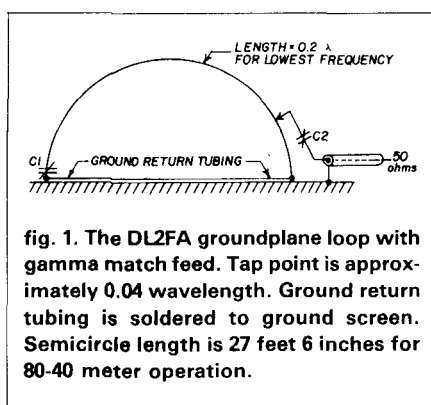


fig. 1. The DL2FA groundplane loop with gamma match feed. Tap point is approximately 0.04 wavelength. Ground return tubing is soldered to ground screen. Semicircle length is 27 feet 6 inches for 80-40 meter operation.

tubing that runs along the surface of the ground is soldered to a groundplane or mat made of chicken-wire (size not specified).

Measured data for the groundplane loop is tabulated in **table 1**.

FIBS revisited

A few years ago 80-meter operators in southern California and Arizona were surprised to hear one of the rarest broadcasting stations in the world: the *Falkland Islands Broadcasting Station* operating on 3.958 MHz. The signals would appear on an otherwise dead band in the late afternoon hours for a few moments and then disappear. The signals seemed to peak in the spring and fall.

Reports indicate the station is being heard once again. It seems that they are using single sideband with carrier, with the audio present on the

upper sideband. Reception reports on this unusual station would be appreciated. The programs consist mostly of pop and disco music with BBC news on occasion.

It is thought that the FIBS propagation into the Northern Hemisphere is by other than normal ionospheric reflection. One theory is that the signal (and other DX signals near that frequency) may be propagated by a "whistler" path which extends into outer space far beyond the ionosphere.

a simple 50 to 75 ohm RF transformer

Shown in **fig. 2** is a simple high-frequency transformer that provides wideband transformation between 50 and 75 ohm unbalanced lines. The design is by JA1AKP, and appeared in a recent issue of *CQ-ham radio* (Japan). The transformer is wound on a FT-114 ($\mu = 125$) No. 61 ferrite core (Amidon FT-114-61) having an outside diameter of 1.142 inches and an inside diameter of 0.75 inch. A trifilar winding of No. 18 wire is used, with two coils having 8 turns and one coil having 4 turns.

A 100 pF mica capacitor to ground is placed at the junction of the two 8 turn coils and a 10 pF capacitor to ground is placed at the 75 ohm port. The units are 500 volt rated. The balun is designed for 3.5 to 30 MHz at the 100 watt average power level.

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ancient modulation

Since the late 1950's single sideband has reigned supreme on the Amateur bands. Users of amplitude modulation ("ancient modulation") have been greeted with derision and have gradually retreated to obscure regions of 160 and 10 meters. Many newly-licensed Amateurs have never heard an AM ham signal!

While Amateurs may pride themselves that sideband has provided them with a quantum leap ahead in communications reliability, they may forget that amplitude modulation is still the prime communication medium in the world. After all, the broadcast band is full of AM signals, and there are a lot of them!

Generally speaking, while the quality of Amateur SSB voice transmission has deteriorated,* the quality of AM transmission has improved. The modern term for broadcast quality is "transparency" — the transmitting system must be "transparent" with regards to the modulating signal.

While much thought and ingenuity have gone into AM transmission in the last decade, most of it has bypassed the Radio Amateur. New techniques, such as high-level pulse-duration modulation, AM stereo and audio processing are unknown in the ham bands, but noise reduction systems, equalizers, delay/reverb effects, level controllers, low frequency extenders and other processing techniques are routinely used by AM broadcasters to make their audio more realistic and outstanding.

On the receiving end of the line,

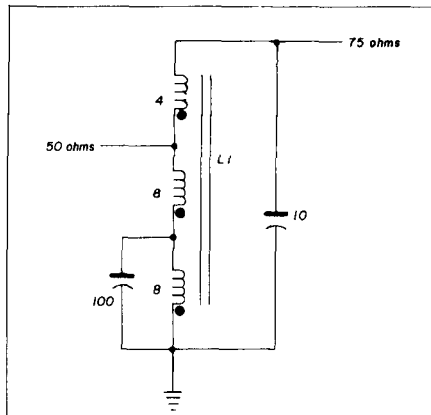


fig. 2. Wideband transformer to match 50 ohms to 75 ohms over the range of 3.5 to 30 MHz. Amidon FT-114-61 (or equivalent) core is used. Trifilar winding of No. 18 wire consists of two coils of 8 turns and one coil of 4 turns. See text for further details.

even the less expensive clock-radio combo receivers have audio systems that make the average ham receiver blush with shame.

Even though most ham signals are "strained" through a narrow-band IF filter, there is no reason why the rest of the receiver should be below par as far as audio quality goes.

Recently I put a little audio feedback around the output stages of my trusty ham receiver, replaced the 19-cent (?) output transformer with one that had some iron in it, and ran the whole works into a spare stereo speaker. My goodness! I didn't recognize the sound of the ham bands! A lot of signals (whose operators previously sounded as if they were talking into a tin can at the end of a

long string) actually sounded quite pleasant, and some signals sounded downright natural!

I then wondered what my transmitted audio quality was like. I put a good audio generator into the transmitter and looked at the audio signal as it entered the balanced modulator. I didn't like what I saw. Any resemblance to a sine wave was minimal.

It was no job at all to add a feedback resistor and to change another resistor in the low-level audio stages to make the waveform look very passable. Now, what came out of the exciter audio system bore a close similarity to what went in.

At this stage I stopped as I became bogged down into an interminable discussion of microphones with friends who knew a lot more about them than I did. The consensus was that the best way to improve audio quality was to throw away the microphone that came with the exciter and buy a good one! Without exception, all of my friends who had particularly pleasing and "punchy" signals on the air had replaced their standard transmitter microphones with different (and more expensive) units.

So there it is. The "ancient modulation" techniques of 1984 and the techniques used by up-to-date broadcasters certainly have some merit for use in the Amateur bands. Which of these new techniques should we examine? I throw the column open to discussion from the readers. (What's the input from hams who work in broadcasting? I'd especially like to hear from you.)

staggered stacking

It is common practice in the VHF bands to stack two or more high-gain Yagi beams to obtain increased gain. Adjustment of the stacking distance can improve sidelobe reduction of the array.

Experiments conducted by G.J. McDonald, VK2ZAB, and reported in the Australian magazine *Amateur Radio*, indicate that an improved

table 1. Groundplane loop dimensions and performance data. (Courtesy of *Radio Communication*, published by the Radio Society of Great Britain and available from Ham Radio's Bookstore, Greenville, New Hampshire 03048.)

band	semicircle length	C1 (pF)	bandwidth (-3 dB points)	efficiency % (MHz)	comparison with 1/4 wave vertical
3.5 - 7.0	27 ft 6 in (8.4 m)	332	5.7 kHz	70 (3.5)	-1.91 dB
			to 67 kHz	96 (7.0)	-0.55 dB
7.0 - 14.0	13 ft 9 in (4.2 m)	184	11 kHz	77 (7.0)	-1.52 dB
			to 147 kHz	97 (14.0)	-0.50 dB
14 - 28	6 ft 10 in (2.1 m)	102	24 kHz	83 (14.0)	-1.22 dB
			to 323 kHz	98 (28.0)	-0.47 dB

*See "better-sounding SSB," page 58 — Editor

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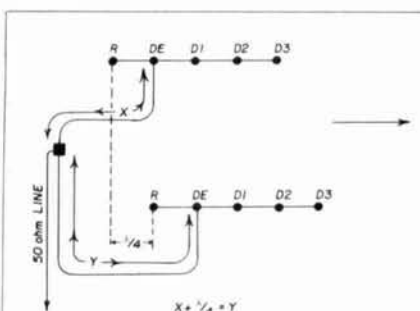


fig. 3. The VK2ZAB array employing staggered stacking. Bottom Yagi is advanced one-quarter wavelength in front of upper array and extra quarter wavelength is added to the phasing line. Front-to-back ratio is improved by this modification.

front-to-back ratio in a Yagi stack can be achieved by staggering the antennas. That is, the bottom Yagi of each vertically stacked pair is advanced one-quarter wavelength on its mounting so that it projects one-quarter wavelength in front of the upper array (fig. 3). The phase lead in the forward direction is cancelled by adding an extra quarter wavelength to the phasing harness connected at the bottom Yagi, thus insuring that the current to both arrays arrives from the feed junction in phase.

Signals arriving at the rear of the array are subject to a 90-degree lag by virtue of the position of the bottom Yagi relative to the top one, plus the additional 90-degree lag caused by the extra length of phasing line.

Results? With a conventional stack of two Yagis, the measured front-to-back ratio was about 12.5 dB. The staggered stack provided a front-to-back ratio of 24 dB, for an improvement of 11.5 dB. And that is equivalent to about 2 "S" units — enough to make the difference between just being able to copy a DX signal through QRM from the rear of the array and copying it comfortably!

Final measurements showed the staggered stack had about 0.25 dB more gain than the conventional stack and also reduced the sidelobes to some extent.

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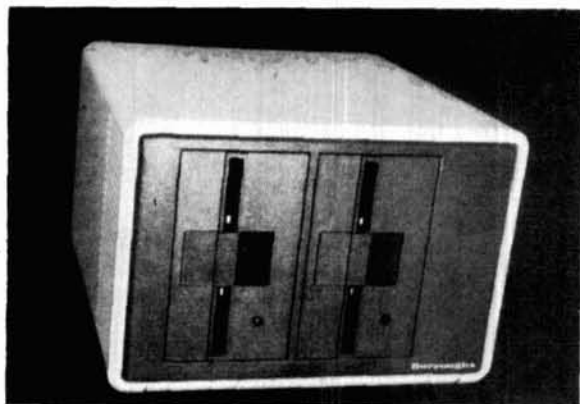
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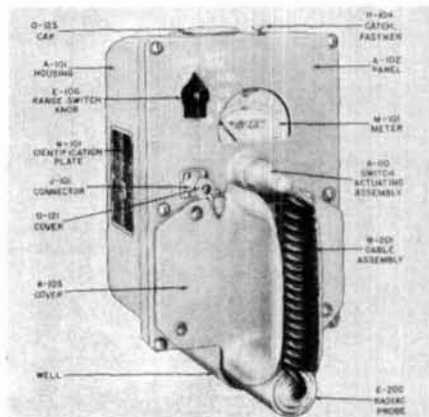
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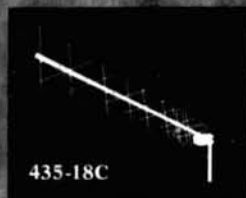
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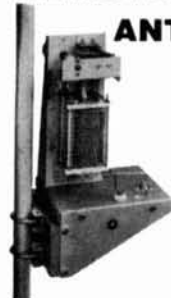
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VHF meteor scatter communications

Sporadic meteor activity improves reliability

Meteors are space junk — typically, small bits of stone or metal from a passing comet, or perhaps leftovers from the formation of our solar system.

When a meteor enters earth's atmosphere it almost always burns up. Its violent passage produces heat, light, and an ionized trail of gases in the upper atmosphere which reflect or reradiate radio signals, an effect known by those who observed the return of American astronauts from space. Not long ago, the crews of the space shuttle *Columbia* lost contact with the ground for several minutes during reentry because of the reflective properties of ionized gases created by the tremendous heat of reentry.

A few Amateurs have taken advantage of these ionized trails to communicate beyond the VHF horizon, using various modes of transmission. Until now, however, communications using the reflective and refractive qualities of meteor trails have been restricted to periods of stream or shower meteors, because only these periods provide a level of ionization high enough to support two-way communication using Morse code or voice modes.

Once again the Amateur community stands on a technological frontier, for with the aid of today's microprocessor systems, some of the more than 100 million meteor trails that occur daily can be used

even though their peak reflective trail lasts only several milliseconds.

Perhaps the most appealing aspect of the meteor-burst concept is the use of bands "closed" to normal communication because of poor conditions. Sporadic meteors occur 24 hours a day, 365 days a year. This means you need not store that 6-meter rig in the closet during minimal sunspot activity; you can go right on using it to communicate in a new way using new transmission techniques.

shower and sporadic meteors

Ionized paths created by the passage of sporadic and shower meteors allow two-way radio communication by returning a portion of the signal to the earth much the same as the ionosphere does at frequencies below 30 MHz. Current Amateur meteor-scatter operations take place mainly during large meteor showers. These predictable showers are composed of meteors large enough to cause trails lasting several minutes; the trails are long enough to allow operating techniques to take place at humanly controllable speeds.

Meteor showers or *stream meteors* occur at predictable times during the year (see **table 1** and **fig. 1**), with peak activity during June, July, August, and December. Communication using the shower meteors is a VHF speciality in the same class as using sporadic E, tropospheric scatter, and ducting conditions. (In contrast, the sporadic meteor-scatter user need not be concerned with meteor-shower schedules or optimum direction calculations.)

Because they appear to radiate from certain points

By Lloyd Vancil, AI7J, 520 Del Sur Way, Oxnard, California 93030

in the sky, meteor showers are usually named for a constellation visible nearby: for example, the Geminids appear to come to Earth from the constellation Gemini. As the constellation moves across the sky, communication is established and maintained by keeping the antenna trained on Gemini as the Earth rotates.

Unlike shower-borne stream meteors, *sporadic meteors* bombard the earth's atmosphere constantly. An average of 100 million meteors large enough to leave an ionized trail enter the earth's atmosphere

daily. The usable trails occur at an altitude of 50 to 75 miles (80 to 120 km) and must last more than a few hundred milliseconds (**table 2**) in order to be usable.

Sporadic meteors are the small meteors characterized by the "ping" noise heard by VHF DXers — hence the name, "ping jockey." Sporadic meteors are the ones used to communicate over distances up to about 1000 miles (1600 km) by relatively new data acquisition systems.¹

Although sporadic meteors appear continuously, their most intense activity is between 0400 and 0800

table 1. Meteor shower data for 40 degrees north latitude. Peak days, meteor velocities, and hourly rates are given, in that order, under shower name.

shower and date (peak date) (velocity) (rate)	optimum time/path (local standard)	antenna offset direction	notes
January 1-4 QUADRANTIDS Jan. 3, 42km/s 50/hr	0220-0740/NW-SE 0740-0900/E-W 0900-1430/NE-SW	SW S SE	intense, but lasts only several hours
May 1-6 ETA AQUARIDS May 4, 64km/s 15/hr	0410-0630/NE-SW 0630-0830/E-W 0830-1050/NW-SE	NW N NE	
June 1-15 ARIETIDS June 7-8, 39km/s 60/hr	0515-0745/N-S 0745-0920/NE-SW 0920-1015/E-W 1015-1155/NW-SE 1155-1425/N-S	W NW N NE E	intense daylight shower, but mostly small particles
July 26-31 DELTA AQUARIDS July 29-30, 43km/s 20/hr	0100 /NE-SW 0100-0330/E-W 0330 /NW-SE	NW N NE	
August 10-14 PERSEIDS Aug. 12-13, 61km 50/hr	2330-0430/NW-SE 0430-0730/E-W 0730-1230/NE-SW	SW S SE	reliable; many large particles
October 18-23 ORIONIDS Oct. 21-22, 67km/s 20/hr	0015-0145/N-S 0145-0350/NE-SW 0350-0500/E-W 0500-0710/NW-SE 0710-0840/N-S	W NW N NE E	
November 14-18 LEONIDS Nov. 16-17, 72km/s 10/hr	0200-0430/N-S 0430-0600/NE-SW 0600-0700/E-W 0700-0840/NW-SE 0840-1100/N-S	W NW N NE E	Periodic, spectacular every 33 yrs; poor in between. Next peak in 1999.
December 10-14 GEMINIDS Dec. 12-13, 35km/s 60/hr	2115-0100/N-S 0100-0200/NE-SW 0200-0230/E-W 0230-0330/NW-SE 0330-0715/N-S	W NW N NE E	reliable, many large particles
December 21-23 URSIDS Dec. 22, 35km/s 13/hr	2200-1600/E-W	S	

hours, local time, because this is the time when the local sector turns toward the direction of the Earth's travel through space. This is a solar system event, in that the occurrence is approximately the same at local time at any point on the Earth — just as daybreak or moonrise.

The peak in daily variation in sporadic meteor activity occurs when a sector of the Earth's surface becomes the "leading edge" of the planet's progress through space, effectively "leading" the sweep of the planet through space.

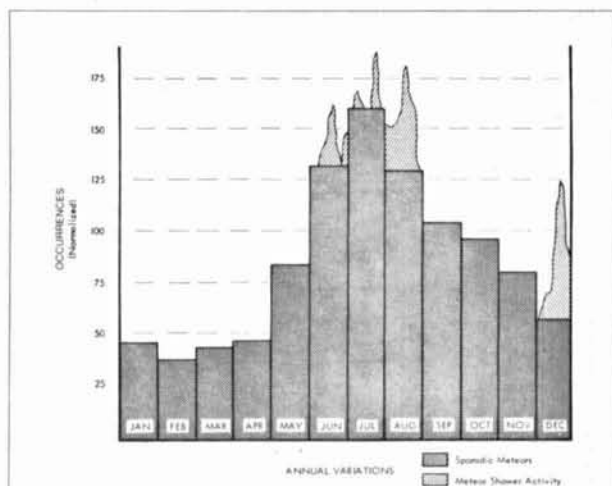


fig. 1. Normalized meteor activity by month.

From late afternoon through early evening, the "leading-edge" portion of Earth's surface changes to the "trailing edge;" only those meteors traveling fast enough to catch up are caught in Earth's gravitational field. Of these, only some will produce a usable trail. The late afternoon events are much further apart and meteors move much more slowly relative to the Earth's surface, resulting in greatly increased waiting time between usable paths (fig. 2).

Because the number of usable sporadic trails varies on an annual and daily basis, astronomers have concluded that sporadic-meteor materials are spread fairly evenly throughout the solar system, but confined to the plane that intersects the planets.

history

Mankind has been aware of meteors — so called "shooting stars" — since his first view of the night sky. The radio disturbance caused by meteors is a relatively recent discovery; Skellett,² in 1932, was the first to link an increase of noise in a receiver to the passage of meteors through the atmosphere. Allen³ studied the meteor-scatter propagation of signals from an FM transmitter at Paxton, Massachusetts, in 1934. In this work, he identified both the shower and sporadic types of meteor trails. Virtually all advances in the field to date are based on his and Skellett's works.

The feasibility of meteor-burst communications was first tested by the National Bureau of Standards

table 2. Order of magnitude estimates of the properties of sporadic meteors.

	mass (grams)	radius	number of this mass or greater swept up by the earth each day	electron line density (electrons per meter of trail length)
particles pass through the atmosphere and fall to ground	10 ⁴	3.15in (8 cm)	10	10 ¹⁸
Particles totally disintegrated in the upper atmosphere	10 ³	1.57in (4 cm)	10 ²	10 ¹⁸
	10 ²	7.87 × 10 ⁻¹ (2 cm)	10 ²	10 ¹⁸
	10	3.15 × 10 ⁻¹ (0.8 cm)	10 ⁴	10 ¹⁸
	1	1.57 × 10 ⁻¹ (0.4 cm)	10 ⁵	10 ¹⁷
	10 ⁻¹	7.87 × 10 ⁻² (0.2 cm)	10 ⁶	10 ¹⁶
	10 ⁻²	3.15 × 10 ⁻² (0.08 cm)	10 ⁷	10 ¹⁵
	10 ⁻³	1.57 × 10 ⁻² (0.04 cm)	10 ⁸	10 ¹⁴
	10 ⁻⁴	7.87 × 10 ⁻³ (0.02 cm)	10 ⁹	10 ¹³
	10 ⁻⁵	3.15 × 10 ⁻³ (80 MICRONS)	10 ¹⁰	10 ¹²
Approximate limit of radar	10 ⁻⁶	1.57 × 10 ⁻³ (40 MICRONS)	10 ¹¹	10 ¹¹
	10 ⁻⁷	0.787 × 10 ⁻³ (20 MICRONS)	10 ¹²	10 ¹⁰
measurements	10 ⁻⁸	0.315 × 10 ⁻³ (8 MICRONS)	?	?

Note: Information from K. Davies, "Ionospheric Radio Propagation," National Bureau of Standards Monograph #80, April, 1965.

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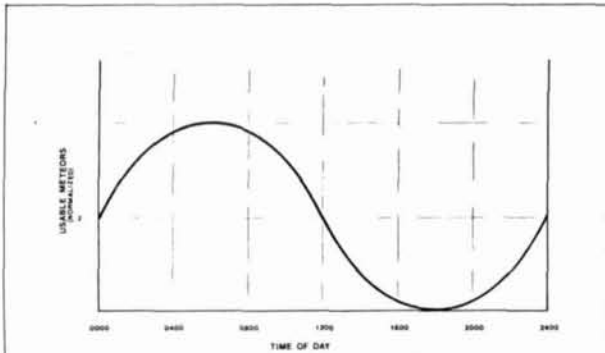


fig. 2. Normalized daily variations in meteor activity.

(NBS) between Cedar Rapids, Iowa, and Sterling, Virginia, a distance of 733 miles (1180 km).⁴ A system built by NBS and operated from 1950 to 1952 showed the feasibility of long-range meteor-scatter communication in the 30- to 100-MHz band. Through experience with this system, NBS defined certain operational characteristics of meteor scatter.

Stanford Research Institute (SRI)⁵ built the first true meteor-scatter communication system in 1956-57 between Bozeman, Montana, and Palo Alto, California. The SRI system used a two-frequency scheme with a burst rate of 454 bits per second (BPS). A pilot frequency containing no information was continuously transmitted. When the receiving station sensed the pilot signal, it would order the system to begin transfer of data. There were several problems with this approach; mechanical delays in the start-up of data transmission and low receiver sensitivity considerably limited the data transfer rate.

In 1956, the Canadian Defense Research Board⁶ began using a system called JANET to conduct experiments over paths of 560 to 745 miles (900 to 1200 km). JANET showed a wide difference in hourly data transfer rates because of daily changes in the number of meteors per unit time and other ionospheric effects, such as sporadic E and aurora. JANET's overall average information rate was 25.5 BPS, which is equivalent to a speed of 38.25 words per minute (WPM).

The National Bureau of Standards resumed meteor communication experiments in 1959, combining JANET technology with some other innovations to attain an overall average data rate of 45 WPM with an error rate of 0.35 percent.⁷ The same year, under contract with the United States Air Force, Hughes Aircraft Company developed a system for air-to-ground communication, introducing techniques for improved block transmission and error detection and correction.

In 1965, the North Atlantic Treaty Organization began operating a system called COMET (COmmunication by MEteor Trails) over a 625-mile (1000-km) path using a 200 watt transmitter and a system for Automatic Repeat Request (ARQ).⁸ ARQ is a system of error detection that allows the receiving station to



fig. 3. Approximate data collection coverage of SNOTEL system.

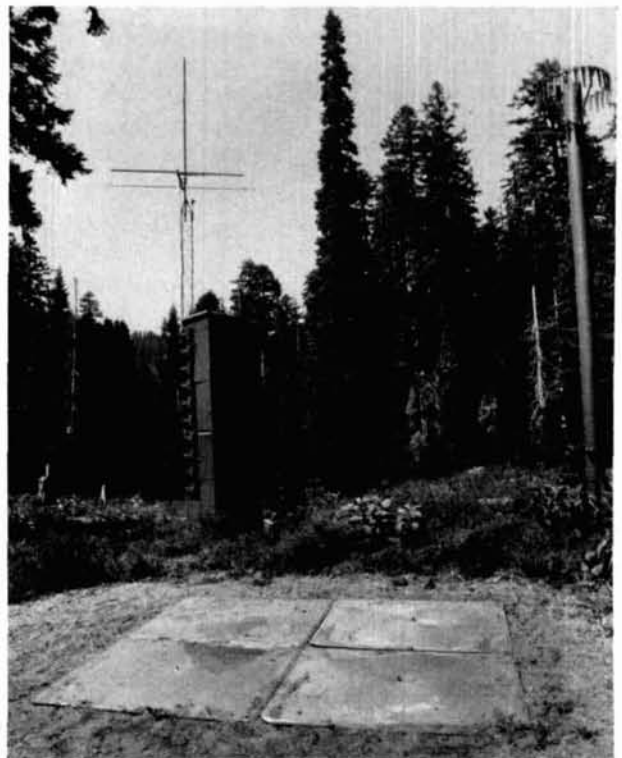


fig. 4. SNOTEL remote site with snow water sensors in foreground.



fig. 5. SNOTEL remote site in winter.

automatically request a repeat of the last information received. COMET achieves a worstcase data rate of 14 WPM and an average of 140 WPM.

largest meteor-scatter link

The Soil Conservation Service of the U.S. Department of Agriculture operates the world's largest meteor-scatter communications data collection system (fig. 3) installed over a period of four years (1976-1980), it is a hydrometeorological data collection system called SNOTEL (*SNO*w *TELE*metry) located in the western United States.

Located in isolated mountain areas, the nearly 500 SNOTEL installations (figs. 4 and 5) collect and transmit detailed data on weather and snow conditions, updating information every 15 minutes. The data is used by the Soil Conservation Service for a number of purposes, including forecasting the amount of water available for irrigation, municipal water supplies, and power generation the following spring and summer.

SNOTEL is controlled by a computer in Portland, Oregon; the two master stations in Ogden and Boise (fig. 6) can request, receive, and acknowledge data from any of the remote sites, using phase shift keyed transmitters with Henry "4K ULTRA" finals and four phase locked loop-based receivers. Each of the four antennas at the master station is a five-element Yagi directed at a different quadrant of the sky and tipped up 30 degrees to optimize performance (fig. 7). Currently, the system collects data from a high percentage (typically 90 to 95 percent) of the more than 1015 active channels in the system during a 2-hour nominal polling period each day — a collection of 7000 characters or the equivalent of 1400 five-letter words per hour, and in maximum message lengths of 256 to 512 bits of sensor data.

possible Amateur applications

Current microprocessor technology opens up the field of meteor burst for the Amateur Radio community. An Amateur meteor-scatter system can be constructed using a microprocessor to control the receive and transmit activity. Literally *anything* that can be digitized — pictures, RTTY data — can be transmitted via meteor burst.

The key to an Amateur meteor-burst communication system is a standardized format, bit rate, and an "acknowledge" or ARQ system. Several versions of ARQ system introduced by current meteor-scatter technology are adaptable to Amateur Radio use. Microprocessor technology offers several schemes for developing ARQ systems for error detection and transmission requests.



fig. 6. Interior of master station at Boise, Idaho.

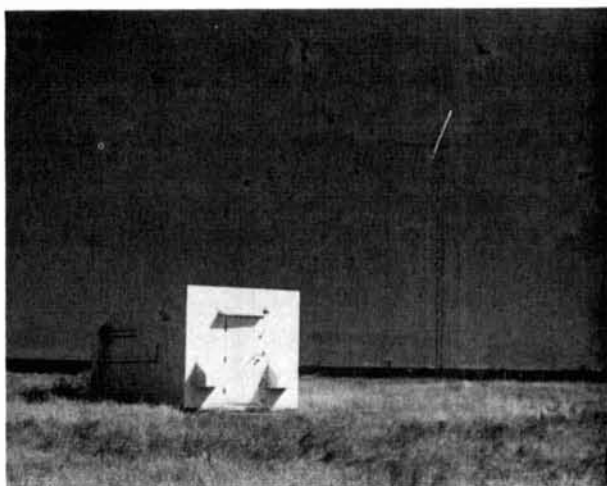


fig. 7. SNOTEL master station located at Boise, Idaho, and one of its four antennas.

A workable format, detailed in fig. 8, consists of 16-bit sync train (two 8-bit bytes), 72 words of data (characters), one byte of parity information, and one group number byte. (This is similar to the SNOTEL data stream and is convenient for computer use because of the 72 characters per line.) Inclusion of the group number allows the receiving end of the communications link to request repeat of a single burst or group of bursts. Basically this is a simple extension of the computer-monitored RTTY station that grew out of the Auto-Start-Auto-Answer RTTY station.

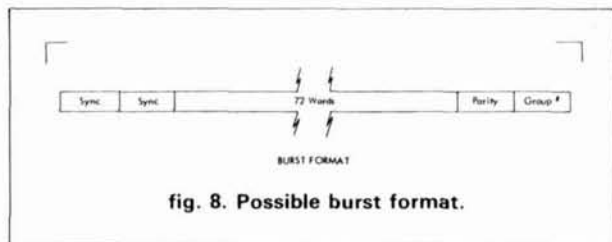


fig. 8. Possible burst format.

conclusion

Almost unlimited opportunity for discovery exists in the field of meteor burst communications. Areas open to Amateur experimentation include antenna design and direction control, error detection and correction techniques, and frequency dependence. Amateurs could take an active role in the improvement of remote control and RTTY systems and in perfecting techniques of data collection and monitoring.

With their access to technology appropriate for use in exploration of the field, Amateur Radio operators enjoy a unique opportunity for innovation and exploration in this expanding field.

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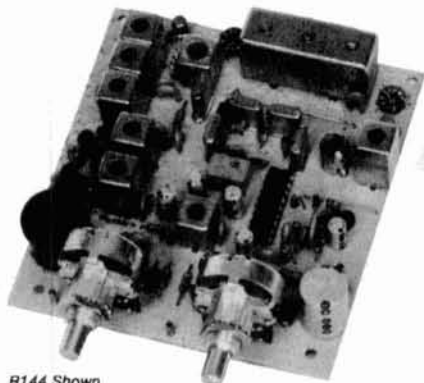
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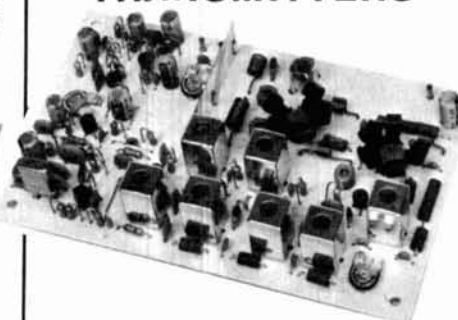
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- **AUTOPATCH KITS.** Provide repeater autopatch, reverse patch, phone line remote control of repeater, secondary control via repeater receiver. Many other features. Only \$90. Requires DTMF Module.



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 - HRF-220 for 213-233 MHz \$38
 - HRF-432 for 420-450 MHz \$48

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Models LNA(), P30, and P432 shown

Model	Tunable Freq Range	Noise Figure	Gain	Price
LNA 28	20-40	0.9 dB	20 dB	\$39
LNA 50	40-70	0.9 dB	20 dB	\$39
LNA 144	120-180	1.0 dB	18 dB	\$39
LNA 220	180-250	1.0 dB	17 dB	\$39
LNA 432	380-470	1.0 dB	18 dB	\$45
LNA 800	470-960	1.2dB	15 dB	\$45

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- P30K, VHF Kit less case \$18
- P30W, VHF Wired/Tested \$33
- P432K, UHF Kit less case \$21
- P432W, UHF Wired/Tested \$36

P432 also available in broadband version to cover 20-650 MHz without tuning. Same price as P432; add "B" to model #.

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Our lab has developed a new line of low-noise receiver preamps with helical resonator filters built in. The combination of a low noise amplifier similar to the LNA series and the sharp selectivity of a 3 or 4 section helical resonator provides increased sensitivity while reducing intermod and cross-band interference in critical applications. See selectivity curves at right. Noise figure = 1 to 1.2 dB. Gain = 12 to 15 dB.

Model	Tuning Range	Price
HRA-144	143-150 MHz	\$49
HRA-220	213-233 MHz	\$49
HRA-432	420-450 MHz	\$59
HRA-()	150-174MHz	\$69
HRA-()	450-470 MHz	\$79



Models to cover every practical rf & if range to listen to SSB, FM, ATV, etc. NF = 2 dB or less.

	Antenna Input Range	Receiver Output
VHF MODELS	28-32	144-148
Kit with Case \$49	50-52	28-30
Less Case \$39	50-54	144-148
Wired \$69	144-146	28-30
	145-147	28-30
	144-144.4	27-27.4
	146-148	28-30
	144-148	50-54
	220-222	28-30
	220-224	144-148
	222-226	144-148
	220-224	50-54
	222-224	28-30

	432-434	28-30
UHF MODELS	435-437	28-30
Kit with Case \$59	432-436	144-148
Less Case \$49	432-436	50-54
Wired \$75	439.25	61.25

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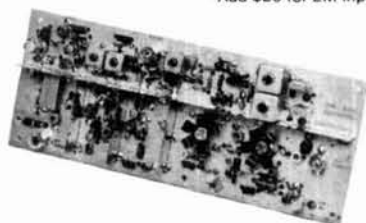
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	Exciter Input Range	Antenna Output
For VHF,	28-30	144-146
Model XV2	28-29	145-146
Kit \$79	28-30	50-52
Wired \$149	27-27.4	144-144.4
(Specify band)	28-30	220-222*
	50-54	220-224
	144-146	50-52
	50-54	144-148
	144-146	28-30

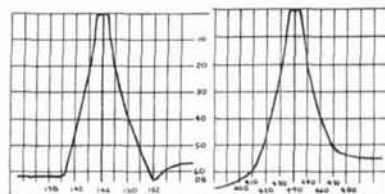
	28-30	432-434
For UHF,	28-30	435-437
Model XV4	50-54	432-436
Kit \$99	61.25	439.25
Wired \$169	144-148	432-436*

*Add \$20 for 2M input



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- 6 pc. AAA Batteries
- DC Cable
- Instruction sheet

ACCESSORIES AVAILABLE

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|---|---------|
| ■ MX Channel crystal.... (Standard Frequency) | \$7.00 |
| ■ MS-1 External Speaker-Microphone | \$23.50 |
| ■ Noise Blanker Kit | \$6.50 |
| ■ NB-1 Side Tone Kit | \$11.50 |
| ■ SP-15 Telescoping antenna | \$19.50 |
| ■ 2M2 DC-DC Converter set | \$17.50 |
| ■ PR-1 Mobile Rack Kit | \$23.50 |
| ■ VX-15 External VXO (one crystal supplied) | \$53.50 |
| ■ PL-15 10W Linear amplifier | \$89.50 |



Photo shown MX-15, VX-15, PL-15, SP-15, MS-1 and PR-1

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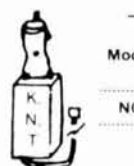
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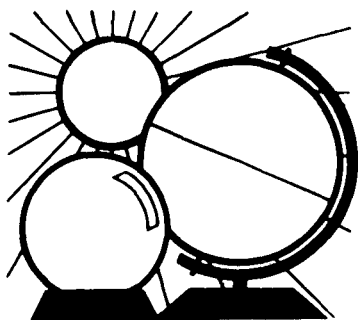
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DX FORECASTER

Garth Stonehocker, KØRYW

last-minute forecast

Excellent DX conditions on the higher frequency bands (10-30 meters) are expected during the first and second weeks of the month due to a probable increased solar flux. A recurrent disturbed period is also possible around February 4, which should result in better transequatorial openings. Check with WWV's geophysical announcement at 18 minutes after each hour for confirmation of the high solar flux and geomagnetic A values.

Other periods of disturbance may be on the 14th and from the 22nd through the 28th. The lower bands should be very good all month and excellent during low solar flux periods in the third week of the month. Look for DX openings to unusual east-west locations during the disturbed days.

No significant meteor showers occur during February. The full moon and lunar perigee are on the 17th.

low band DX

Since publication of the nighttime bands in the propagation chart was begun, have you noticed how high those bands are, even at their diurnal minimum? DX stands for long distance contacts; even the shortest DX path, Europe, is 3800 miles (6200 km) from the East Coast. South America, Japan, and the Far East are approximately 6200 miles (10,000 km), while Antarctica and Australia are 9000 miles (14,600 km) and 9900 miles (16,000 km), respectively. The highest usable bands are related to the maximum usable frequencies (MUF) for that path. The longer DX paths

are to the south across the equator where minimum MUFs are around 14-15 MHz now and will decrease by 3 MHz when the sunspot number drops to 16 (or the radio flux to 75 units). The other DX paths are in mid to high latitudes where MUFs are lower and paths are shorter, giving minimum MUFs of only 8-9 MHz. These will decrease by 2 MHz at the sunspot cycle minimum.

These numbers are based on monthly median (January) data. The daily solar flux factors from last month's *DX forecaster* can be used to forecast several days ahead. The *minimum* MUF is inversely related to the daily flux, decreasing with an increasing change in radio flux. Geomagnetic disturbances will also affect these minimum MUFs by 5 to 10 percent — the high latitude paths negatively and transequatorial paths positively.

Interestingly, even at sunspot minimum, 40 meters will be open to Europe, Japan, and the Far East and 30 meters on transequatorial paths. If you work 80 meters and 160 meters, they will be 50 percent and 75 percent below the MUF even on the short-northern paths. This doesn't mean they won't work; we know they do.

band-by-band summary

Ten and fifteen meters will be open for worldwide DX communication from sunrise until after sunset during the twenty-seven day solar flux maximums. Skip of 2500 miles (4000 km) (or multiples) is possible, and will follow the sun across the earth.

Twenty meters will be open to some area of the world for the entire twenty-four hour period on many days

of the month. The band should peak in all directions just after local sunrise, and again toward the east and south during late evening hours. During hours of darkness the band will peak toward the west in an arc from southwest through northwest, encompassing Pacific areas.

Thirty meters is a daytime and nighttime band. The day portion should be like 20 meters except in that signal strengths may decrease during midday on days having high solar flux values. This band will also be useful well into the night and often through the night. Once again exceptions to this are on nights that follow very high solar flux value days. The problem time is usually the hour or so before dawn (diurnal MUF minimum). The workable distance may be expected to be greater than 80 DX at night and less than 20 during the day.

Forty and eighty meters will be the most usable nighttime DX bands. Most areas of the world will be workable from dusk until sunrise. Hops shorten on these bands to about 2000 miles for 40 meters and 1500 miles for 80 meters, but the number of hops can increase because signal absorption in the ionosphere's D-region is low during the night. The path follows the direction of darkness across the earth, similar to the way in which the higher bands follow the sun.

One-sixty meters will be similar to 80 meters, providing good working conditions for enthusiastic DXers who like to work nighttime and early morning hours, especially at local dawn.

ham radio

WESTERN USA

GMT	PST	N	NE	E	SE	S	SW	W	NW
0000	4:00	20	20	15	10	15	10	10	20
0100	5:00	20	20	15	10	15	10	10	20
0200	6:00	20	20	15	10	15	10	10	20
0300	7:00	20	30	15	10	15	10	10	20
0400	8:00	20	30	15	10	15	10	10	20*
0500	9:00	20	30	20	10	15	10	10	20
0600	10:00	20	30	20	15	15	10	10	20
0700	11:00	20	30	20	15	15	10	15	20
0800	12:00	20	30	20	20	15	15	15	20
0900	1:00	20	30	20	20	20	15	15	30
1000	2:00	30	40	20	20	20	15	20	30
1100	3:00	30	40	20	20	20	20	20	30
1200	4:00	30	40	20	20	20	20	20	30
1300	5:00	30	30	20	20	20	20	20	30
1400	6:00	30	30	15	20	20	20	20	40
1500	7:00	40	20	15	15	15	20	20	40
1600	8:00	40	20	15	15	15	20	30	30
1700	9:00	40	20	15	15	15	15	15	30
1800	10:00	40	20	10	15	15	15	15	20
1900	11:00	40	20	10	15	15	15	15	20
2000	12:00	30	20	10	15	15	10	15	20
2100	1:00	30	20	10	10	15	10	10	20
2200	2:00	30	20	15	10	15	10	10	20
2300	3:00	30	20	15	10	15	10	10	20

MID USA

MST	N	NE	E	SE	S	SW	W	NW
5:00	30	20	15	10	15	10	10	20
6:00	30	20	15	10	15	10	10	20
7:00	30	20	15	10	15	10	10	20
8:00	30	30	15	10	15	10	10	20
9:00	30	30	15	15	15	10	10	20
10:00	30	30	20	15	15	10	10	20
11:00	30	30	20	15	15	10	15	20
12:00	30	30	20	15	15	15	15	20
1:00	30	30	20	20	20	15	15	30
2:00	30	40	20	20	20	15	20	30
3:00	30	40	20	20	20	20	20	30
4:00	20	40	20	20	20	20	20	30
5:00	20	30	20	20	20	20	20	30
6:00	20	30	15	20	20	20	20	40
7:00	20	30	15	15	20	20	20	40
8:00	20	20	15	15	20	20	20	40
9:00	20	20	15	15	15	20	20	40
10:00	30	20	10	15	15	15	15	30
11:00	30	20	10	10	15	15	15	30
12:00	30	20	10	10	15	15	15	20
1:00	30	20	10	10	15	10	15	20
2:00	30	20	10	10	15	10	10	20
3:00	30	20	15	10	15	10	10	20
4:00	40	20	15	10	15	10	10	20

EASTERN USA

CST	EST	N	NE	E	SE	S	SW	W	NW
6:00	7:00	30	20	15*	15	15	10	10	20
7:00	8:00	30	20	15	15	15	10	10	20
8:00	9:00	30	20	15	15	15	10	10	20
9:00	10:00	30	30	15	15	15	10	10	20
10:00	11:00	30	30	15	20	15	10	10	20
11:00	12:00	30	30	15	20	15	10	10	20
12:00	1:00	30	30	15	20	15	10	10	20
1:00	2:00	30	30	20	20	15	15	15	20
2:00	3:00	30	30	20	20	20	15	15	20
3:00	4:00	30	30	20	20	20	20	20	30
4:00	5:00	30	40	20	20	20	20	20	30
5:00	6:00	20	40	20	20	20	20	20	30
6:00	7:00	20	30	15	20	20	20	20	40
7:00	8:00	20	30	15	15	20	20	20	40
8:00	9:00	20	30	15	15	15	20	20	40
9:00	10:00	20	20	15*	15	15	20	20	40
10:00	11:00	20	20	10	15	15	20	20	40
11:00	12:00	20	20	10	15	15	15	15	30
12:00	1:00	20	20	10	15*	15	10	15	20
1:00	2:00	20	20	10	10	15	10	15	20
2:00	3:00	30	20	10	10	15	10	10	20
3:00	4:00	30	20	10	10	15	10	10	20
4:00	5:00	30	20	15	15	15	10	10	20
5:00	6:00	40	20	15	15	15	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during 'normal' hours.
 *Look at next higher band for possible openings.

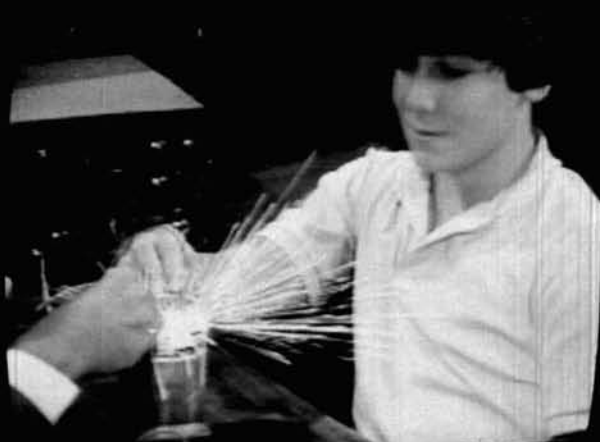
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EMI/RFI shielding: new techniques

part 2

Techniques and materials that reduce EMI

Part 1 of this two-part article served as a basic overview by identifying the problem of EMI/RFI, citing examples of its adverse effects, and providing a general discussion of EMI shielding effectiveness (SE). In one form or another, each of these various schemes involved plastics. This more detailed, more technical article examines methods of testing SE that employ RF techniques familiar to most hams.

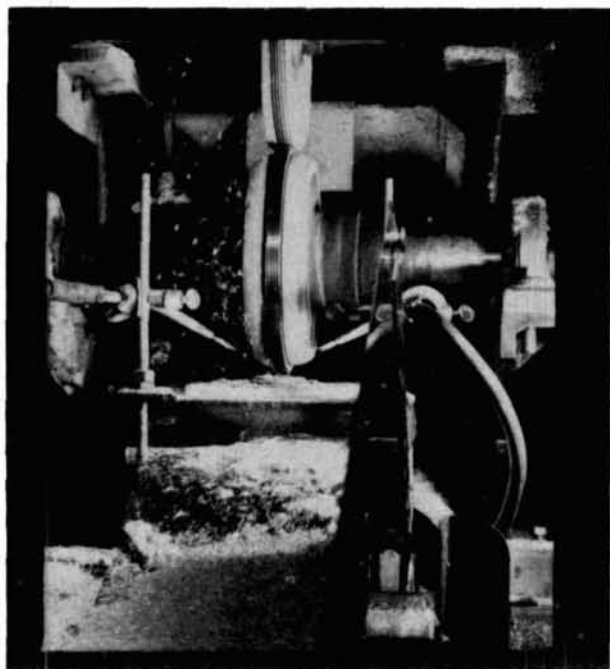


fig. 1. The generation of aluminum flakes.

Before examining SE and its testing, let's first examine the process of fabricating plastics in a manner that renders them effective for EMI shielding and containment, yet is economical enough to allow plastic enclosures to be used instead of metal ones.

EMI and plastics

The FCC tries to stringently enforce EMI emission standards that apply to electronic equipment. The pressures of foreign competition, however, demand that American manufacturers develop innovative cost-cutting manufacturing techniques such as using EMI-proof plastic enclosures instead of metal enclosures. But there are other ways than the use of enclosures to obtain or shield electronic devices from EMI/RFI; we can use metallic tape with adhesive backs (the adhesive is a plastic), neoprene and elastomer gaskets (both plastics), or sprays and paints in conjunction with plastics. In fact, plastics are so common that we sometimes overlook them, and often fail to fully appreciate or understand their versatility.

In order to better understand how plastics are used in potting and casting, in PC board conformal coating, in insulation, high frequency connectors and a myriad of other applications, let's begin by looking at materials used for shielding.

materials for shielding

Most EMI shielding processes in use today have disadvantages: precious metal powders are expensive; conductive carbon comes in but one color — black — zinc metallic sprays can lose adhesion, with small metal flakes falling onto PC boards and shorting out components; and paints are subject to abrasion and subsequent delamination when put through temperature extremes. What is the solution to the problem? The answer may be a new technique of utilizing structured fillers.

By Vaughn D. Martin, 114 Lost Meadows,
Cibolo, Texas 78108

Graphite fillers. With higher conductivity and better reinforcement characteristics than carbon blacks, graphite fillers look promising. Radar chaff, for example, originally developed to resonate at radar frequencies and to produce "phantom" targets for anti-aircraft batteries, has been successfully introduced into these types of composite plastics.

Impregnation of aluminum flakes into plastic is a relatively new process. A spinning wheel throws off flakes of molten aluminum alloy (see **fig. 1**); these flakes, with a tightly controlled aspect ratio (length-to-thickness ratio), are then deposited into a resin which causes them to become oriented in the direction of the flow of the resin. You have no doubt seen fingerlings or small minnows in schools orient themselves upstream in line with the flow of water; these flakes do the same in the resin. This is important because molding or extrusion processes require this. If all flakes are not oriented in the same direction, they experience lateral force and breakage occurs, with a subsequent reduction in their shielding effectiveness.

Typically, aluminum-impregnated plastics contain 18 to 22 percent flakes by volume because electrical

conductivity depends not on the weight ratio of the flakes with their resin "holder," but on the volume ratio of flakes to resin. This means that a low-density polymer like polypropylene would need 46 percent weight, while a high-density polymer like PVC would require only 37 percent flakes to achieve the same 22 percent volume ratio.

the acid test

To date, most testing techniques have been contrived to test conductors or pieces of pure conductive materials. But now, the purely resistive properties of a substance like straight wire no longer exist; the combining of metals with plastics has caused networks to result with both inductive and capacitive reactances.

In **fig. 2**, note that if probes were to pass through the insert on the left at precisely the right spot, they could miss the conductive particles, completely giving a totally insulative test result. But insertion as a layer in the material to the right with its insulating matrix (checkerboard) results in a more meaningful measurement. (Two test methods of determining surface and volume resistivity can be averaged to determine shielding effectiveness from 0.1 to 1000 MHz by the graph in **fig. 3**).

To evaluate how well material acts as a shield, let's first recognize that a pure conductive metal performs its shielding effectiveness by an almost 100 percent *reflection of radiation*, whereas a conductive-composite product reflects about 80 percent of radiation and absorbs 20 percent. Mathematically, the shielding effectiveness, expressed in dB, (SE) + R + A or the sum of the energy Absorbed and Radiated. In actual practice, there is another term, B, added to A and R, but this term describing internal multiple reflections is usually ignored for materials with greater than 10 dB of SE. This would include just about all materials that are even remotely considered to be shielding elements.

test and measurement systems

There are essentially three methods of measurement: the free-space method, which is impractical at lower frequencies; the shielded box method; and the transmission line technique. The shielded box technique, shown in **fig. 4**, is the most popular. This rectangular metal box, made with thick walls and electrically tight seams in the walls, has a transmitting antenna inside it and a sample port (opening) in one wall for radiation emitted through the port. If the specimen to be tested is placed over the port or hole, and the radiated power P_2 measured against the open power level P_1 , SE can be determined by:

$$SE = 10 \log \frac{P_1}{P_2} \text{ (in dB)}$$

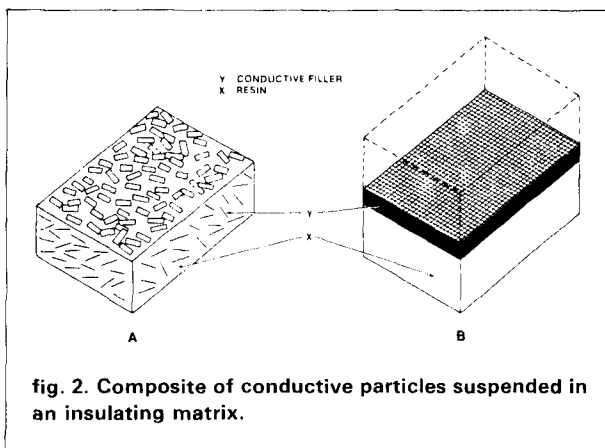


fig. 2. Composite of conductive particles suspended in an insulating matrix.

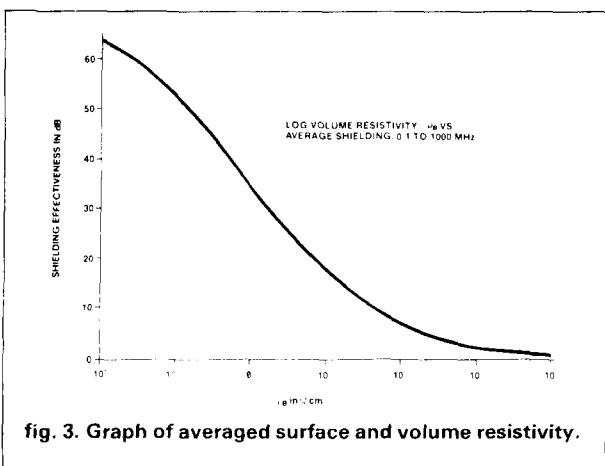
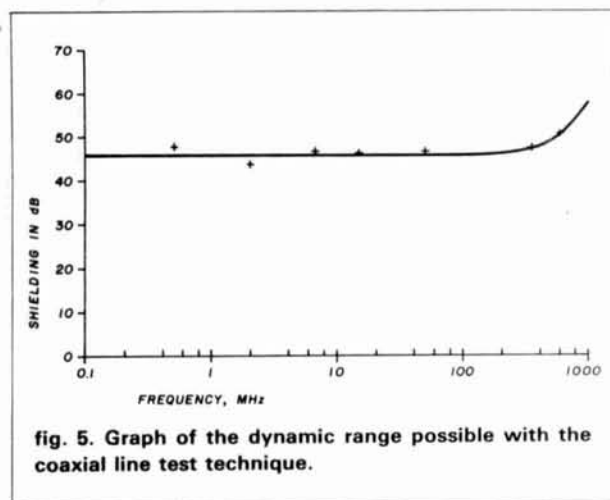
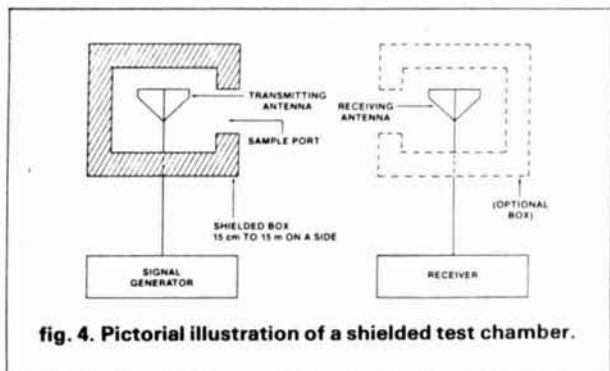


fig. 3. Graph of averaged surface and volume resistivity.



This method of testing has some problems. If there is an edge in the chamber that approaches the wavelength of radiation, resonance can occur with ensuing standing waves between the antennas inside and the chamber's walls. Also, at lower frequencies, the dynamic-range is limited to approximately 50 dB because of the energy that can pass through the small sample port at those wavelengths. It is likely that no two boxes would report the same results for a given sample.

The coaxial transmission line technique has no inherent problems associated with it because this method is not frequency-sensitive. (Note that there are no seams in this canister-shaped specimen holder.) Dynamic range, (fig. 5), is a function of the generator and the sensitivity of the receiver. Unlike the shielded box, which suffered from the wavelength-to-specimen size ratio changing the chamber's own inherent impedance, this problem does not exist with the transmission line technique.

This method (the TEM or transverse electromagnetic cell) yields reflected, transmitted and absorbed data by reflection coefficient measurements, which allows the designer of the system to optimize reflec-

tion. This is ideal for waveguides or parabolic antennas in which the wave is to be redirected.

The main advantage of this measurement technique is the uniformity of repeated test results. The TEM, (fig. 6), allows absolute verification to be made that the reflection and transmission ratios in a 50-ohm line are the same as those theoretically obtained in free space on a panel infinite in extent in the plane perpendicular to the incident energy. The dimensions of a 50-ohm line are such that the ratio of the impedance presented by a test specimen transverse to the direction of propagation, to that presented to the free space wave by the same material, is 0.13. This is the exact ratio of 50/377 (the two characteristic impedances).

Electromagnetic fields near an oscillating dipole antenna have a wave impedance as a function of $1/d$, where d is the distance between the source and receiver. This is true when the distance to the dipole antenna is very small in comparison to the wave-

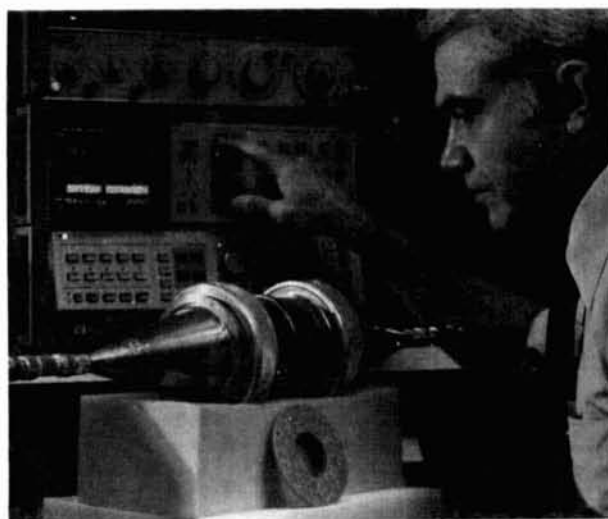
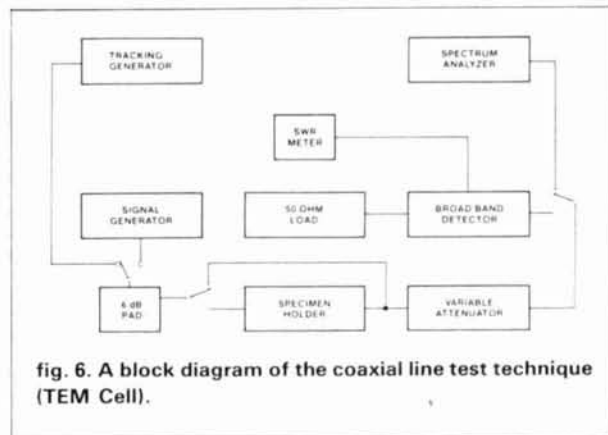
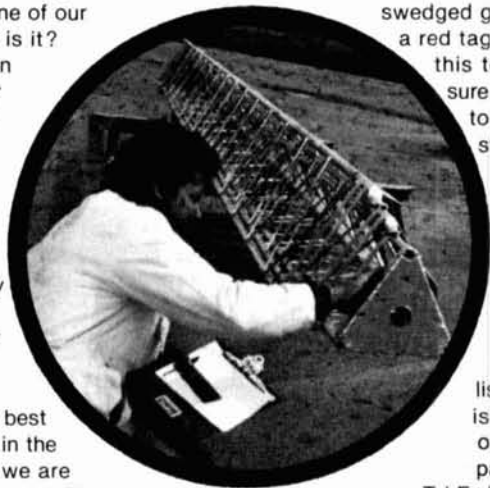


fig. 7. The TEM (Transverse Electromagnetic Cell) test method in operation.

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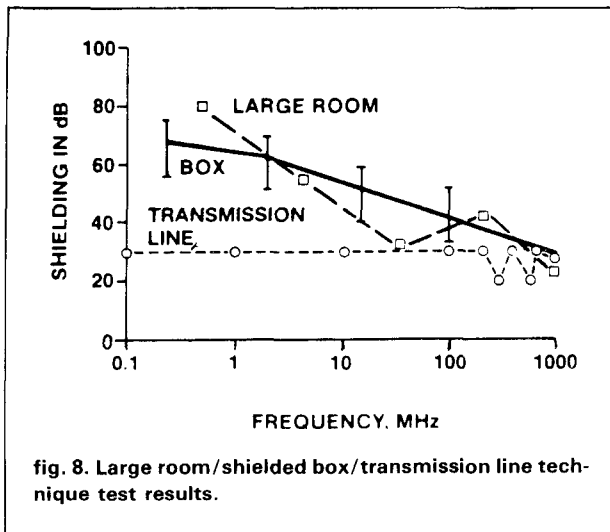
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length and precisely why, at lower frequencies, within a shielded box, the impedance changes with frequency. (We will note some data shortly that graphically reveal this fallacy.) Likewise, as the wave impedance and reflection coefficient increase as frequency decreases, a false indication of an enhanced SE results as the frequency decreases.

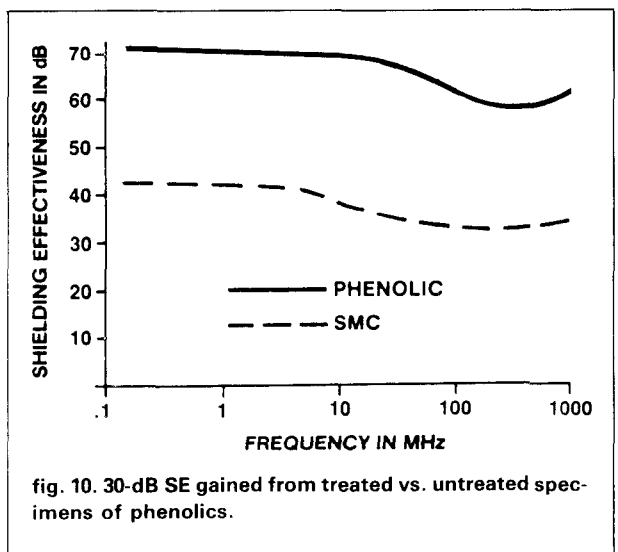
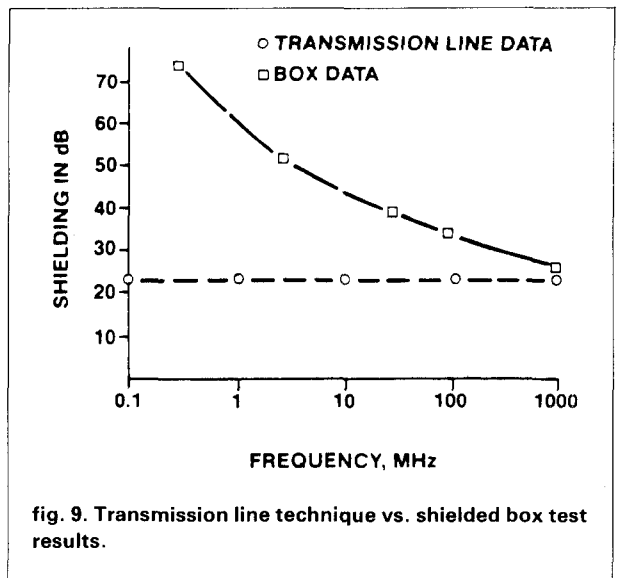
The coaxial transmission line technique is an insertion-loss measurement technique using a substitution method (see fig. 7). The measurements can be made at specific frequencies (the point-to-point mode) using a modulated signal generator, crystal detector, and tuned amplifier (SWR). The other method employs sweeping a band of frequencies with a tracking generator and a spectrum analyzer used as a receiver. The advantage of this point-by-point method is that other information, such as reflection coefficient data, can be taken in this mode; the sweeping mode has the advantage of taking SE data at all points, but without the luxury of allowing observation of reflection coefficients.

In the sweeping mode, the signal generator is replaced by a tracking generator and the variable attenuator removed, because the attenuator in the analyzer is used instead. The spectrum analyzer presents the response over the whole bandwidth in a single curve. However, for determination of SE, two responses are taken, one with and one without the specimen in the holder. The SE is the difference between these two curves.

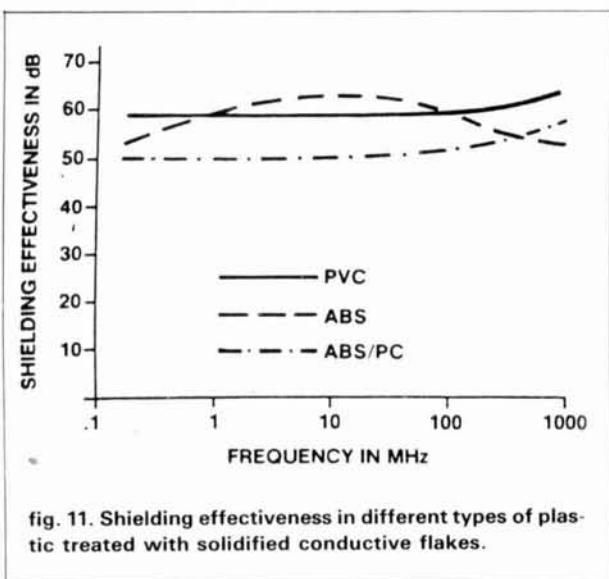
The dynamic range of this measurement technique is determined by three things: the power level of the signal generator, the sensitivity of the receiver, and the degree of shielding of the equipment and the connecting transmission lines. For example, if a one-watt signal generator with a detector such as an HP415 standing-wave meter are used, an 80-dB dy-

amic range can be obtained with standard coaxial cables.

Analysis of test results shows how "flat" the test results are from 0.1 to 1,000 MHz, (figs. 8 and 9). The difference between measured and predicted results is due to the uncertainty of measurement of the conductivity of the composite plastic, as previously discussed. Fig. 8 shows a specimen of 20 percent weight piece of Transmet™ series 100 conductivity modifiers in a polyester resin. The methods used in testing were the shielded box method, using a 6.25 foot (2 meter) cubic box in a 31 × 15.5 × 10 foot (10 × 5 × 3 meter) shielded room; the last method was the transmission line technique. Figs. 10 and 11 show various plastics test results for SE.



*The Transmet Corporation, Columbus, Ohio



the box. These problems are usually of much lower frequency than EMI/RFI emissions and are the result of emissions from motors, transformers, and other magnetic field producers. Most hams know about ferrite beads (fig. 12), small ferro-magnetic metal or graphite tubular-shaped devices that slide over wires and suppress conducted emissions. Ferrite beads, however, become saturated at a certain current level; their attenuation does not increase with frequency; they tend to become a resonant frequency element, and there are often gaps because of poor fittings. Thanks to plastics, there are better ways to suppress conducted emissions. Zippertubing™*, for example, uses metallic mesh, aluminum foil laminated to vinyl-impregnated nylon, and high nickel-steel foil in standard jackets to form a protective shield fit over entire cable bundles (fig. 13), whether they are the traditional round bunched cables or the flat "ribbon" cables used by printers and modern computer equipment today. A new product now under development by Zippertubing, Inc., uses a plastic polyurethane material, PFR-20B, said to be even more rugged than these other jackets.

Berquist** is another innovative company that makes thermally conductive insulating pads that go between resistors and the PC board. This company, however, has gone one step further by laminating copper right onto the pad, (fig. 14). This now provides not only thermal conductivity but also EMI/RFI and conducted interference protection as well. While the metal does cause a capacitance between the transistor and its heatsink, the capacitance is limited to approximately 100 pF for a TO-3 power transistor.

To the ham on a limited budget, the best way to

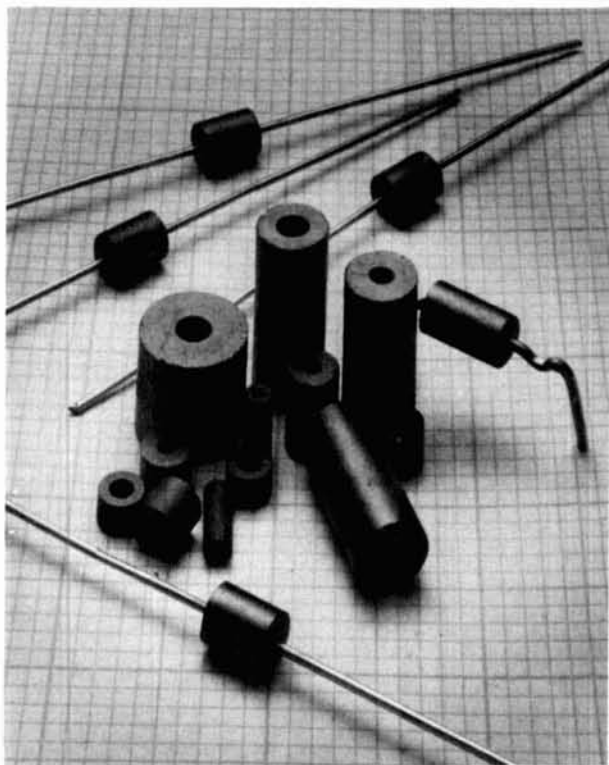


fig. 12. Ferrite beads for EMI suppression.

Although this method of testing is relatively new, it is gaining rapid acceptance. Sometimes called the Stutz method after its inventor, David E. Stutz of the Battelle Institute, this method is used to measure planar waves, whereas the shielded box method is used to measure the near field.

Conducted interference. As stated previously, conducted interference is best attacked at the source or through the cables or wires running to and from



fig. 13. Zippertubing conductive cable covers for suppression of conducted interference.

*Zippertubing, Inc., Los Angeles, California

**Berquist, Inc., Minneapolis, Minnesota

achieve magnetic shielding might be to use foil tape, which is comparatively inexpensive and comes in thicknesses of less than 0.002 inch to over 0.01 inch thick with permeabilities ranging from as low as 3,000 to over 350,000.

(If you are really fascinated by this subject, you may want to spend \$99.50 and buy a kit from the Magnetic Shielding Division of Perfection Mica* (fig. 15). This contains sheets and foil forms, a magnetic pick-up probe, design information, and a variety of different types of shielding.)

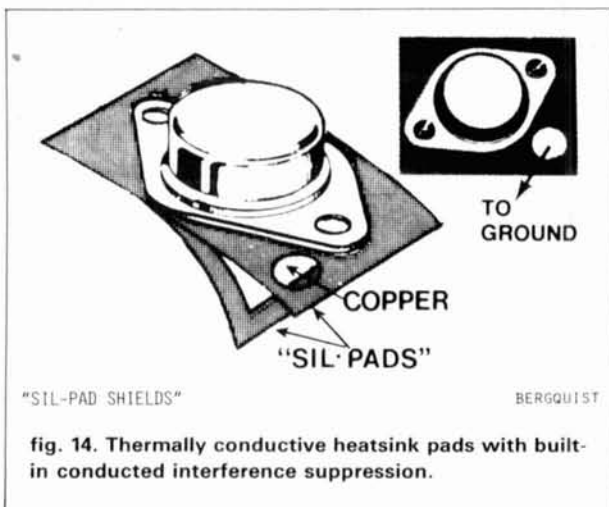


fig. 14. Thermally conductive heatsink pads with built-in conducted interference suppression.

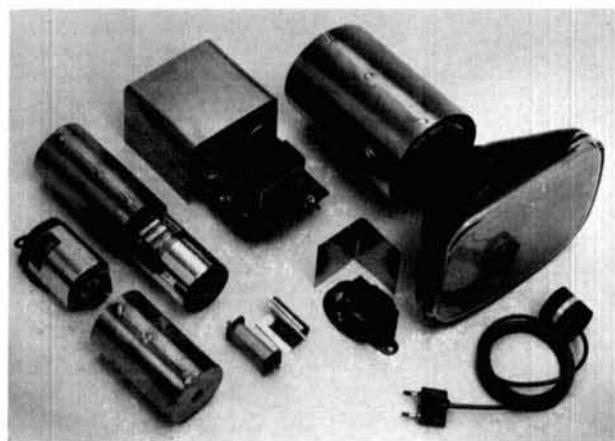


fig. 15. Kit for magnetic field shielding.

There is as yet no perfect coating, adhesive, or potting compound; EMI shielding with plastics is a science in itself. But with technology evolving rapidly in this area — and with many effective EMI-shielding materials becoming increasingly available — you need no longer invariably avoid using plastic cases in building equipment.

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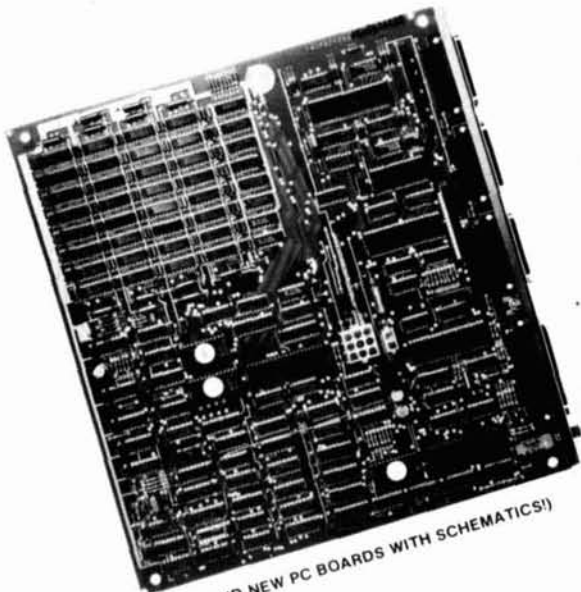
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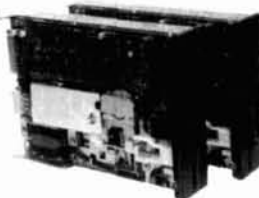
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book and product
REVIEWS

HW-5400 HF transceiver

Regardless of his or her age, every Radio Amateur has heard "the rig here is an HW . . ." Back in the early 1960's, the first of the HW single-banders appeared on the scene, and were an instant hit. That unit spawned a whole line of units which included the SB-101, the SB-301, and many more. There are many more still going strong; and now, there's the HW-5400 microprocessor-controlled HF transceiver.

The HW-5400 has all sorts of interesting features. For instance, when I switched from one band to another, and then back to the first band, it was still tuned to the original frequency. In case of power failure, all the memories are automatically reset to the band edges.

It has split frequency operation, fast/slow AGC, VOX, and antiVOX. It's *all* there. One thing I especially like is the optional keypad frequency entry. I do quite a bit of split frequency work, and this feature makes operating easy; I've also found it handy when going off frequency from a net.

When the kit arrived, the first thing that impressed me was the weight of the power transformer. (Of course, the power supply is rated at 20 amperes, so I shouldn't have been surprised.) As soon as I opened the main carton, I was faced with a notice telling me not to unpack any

parts until I was told to in the assembly procedure. I'm glad I followed the instructions. . . the packaging is indeed dense.

preliminary assembly

The assembly of the kit should take about one hundred hours, more or less. It is NOT a kit for someone who has had no experience in building kits. The instructions and illustrations are up to Heathkit standards, but a good deal of care has to be taken during soldering on some of the more tightly packed boards.

I took slightly less than one hundred hours to do the actual assembly, but I did encounter a couple of problems that used up quite a few extra hours. (It should have taken less, but I was being stubborn and didn't want to ask for help.) One of the areas where I got stuck turned out to be caused by a bad FET. And after some discussions with Heath, it turned out that the company was having some difficulties with this particular part. Heath and Motorola, the manufacturer, were working on the problem. The replacement FETs — which were sent promptly — worked fine, so the problem I encountered appears to have been resolved.

The kit has fifteen circuit boards, some of which are simple, with few parts. Others are not. But all are well made, and the silk screening job is good. While there is really no problem in stuffing these boards, I do recommend that you take the advice of good carpenters: measure twice, cut once.

At the end of the assembly of each board there is a series of tests to be made. In one, I found two solder bridges that I hadn't seen in my visual checks — and I was using a lighted inspection magnifier.

I had a great feeling of accomplishment when all the boards were finally done. Then I came to the section innocently titled "Chassis." At that

moment I didn't realize that I was about to become an expert in crimping and soldering tiny Molex spring connectors on a multitude of ends of wires. After doing some 50 connections, I checked the price of a crimping tool: over \$80. I continued my manual assembly procedures. (Perhaps Heath will develop a design for making a low-cost crimper, because using these connectors really helps during troubleshooting.)

final assembly

As I started the final assembly, I looked at the pile of boards I had just finished, and then at the main chassis, and wondered what kind of magic I was going to have to perform to get all that material onto one tiny chassis. The installation of the pre-assembled main cable harness is a good example of the kind of magic you'll have to perform; I suggest you start this step when *you're feeling well-rested and kindly towards the world.*

After doing the first two or three steps in the final assembly, I was instructed to mount the audio circuit board and make the necessary connections. Then came the instruction to "connect up a power supply" so that a series of operating tests could be made. It was only then I realized I should have built the power supply first. (I think perhaps it would have been wiser of Heath to mention this at the very beginning.) However, fortunately for me, I had "subcontracted" that part of the job to a ham friend who is normally a very active person (he goes up and down 70-foot towers like I go up and down stairs). He'd just had an extensive operation on one knee, and was "climbing the walls" with inactivity. He built my power supply in about six or seven hours.

All went well with the final assembly until the test of the synthesizer board: I just couldn't get the right readings. After following the "in case of trouble" charts and trying a few tricks of my

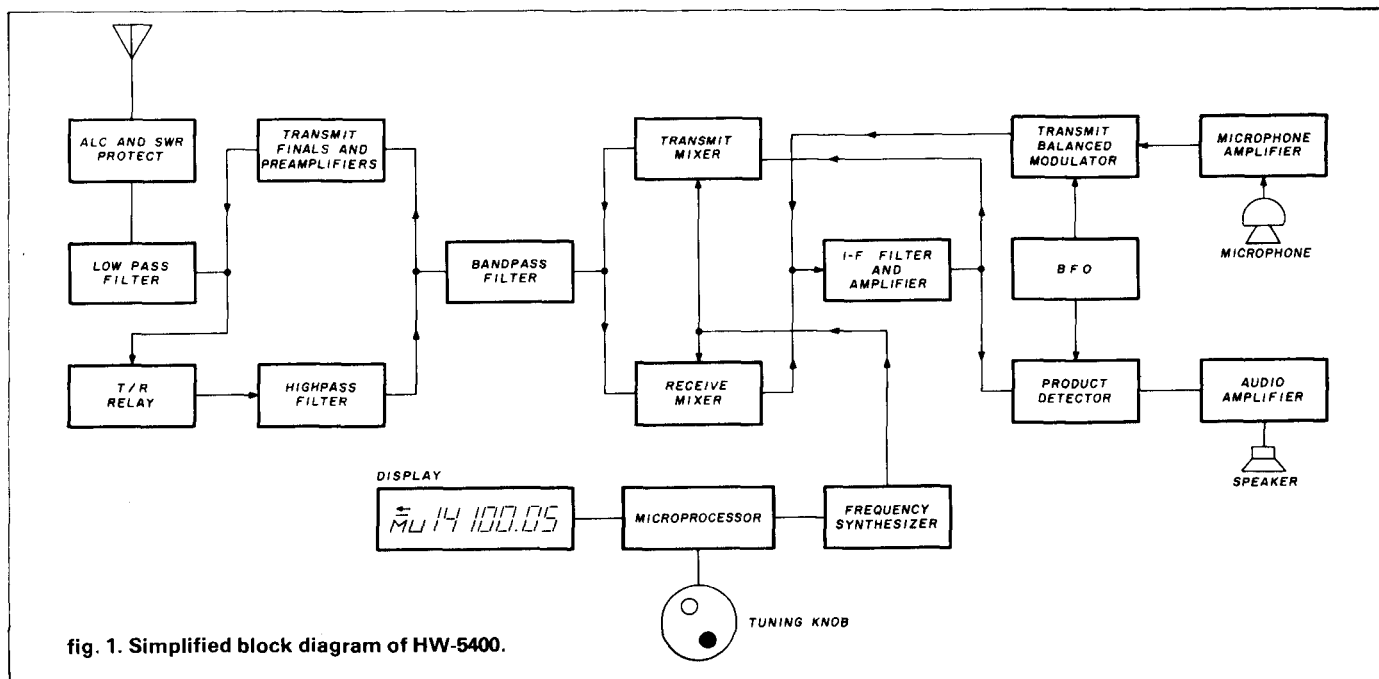


fig. 1. Simplified block diagram of HW-5400.

own, I called Heath. Very shortly, I was talking to Terry, a synthesizer expert, and listening to a quick run-through of that part of the circuit. He made a couple of suggestions and sure enough, his first one identified the trouble. It was a mica tuning capacitor that was marked 100 pF, but was really 1000 pF. No wonder the circuit wouldn't tune!

I ran into difficulty again on the IF board, but this time I didn't waste a moment; I got on the phone. It turned out that the trouble was the FET problem mentioned above. A package arrived in three days, and as it turned out, there were four FETs to be replaced.

When I had to remove the IF board, the synthesizer board, and the RF board, I was quite happy that Heath had made me put all those Molex connectors together. Disassembly was easy. It was at this point that the quality of the printed circuit boards became really obvious. I had absolutely no fear of unsoldering and resoldering the connections on the traces of the PCBs.

"smoke" test

There was one interesting thing about doing power-on tests after each step in the assembly. If you've ever put a kit together, then you know the thrill and fear of the "smoke test." With this rig you enjoy the same emotions over and over again. It doesn't lessen with experience.

Now there it was, all put together and ready for use. After three or four contacts on 75 and 20, I decided it was time to finish the job and install the unit in its cabinet. But first, I thought I should do a bit of peaking up, using my old but reliable General Radio Model 605A signal generator. There was no improvement.

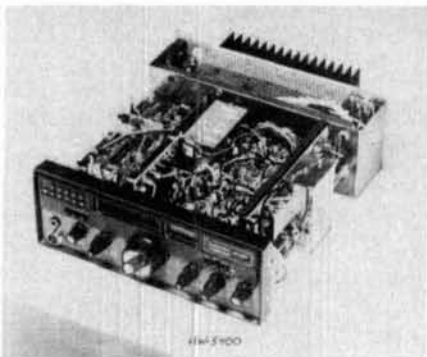
state-of-the-art circuitry

This is a complex, state-of-the-art transceiver. The simplified block diagram (fig. 1) shows the flow of signals. On both transmit and receive, the signals go through the lowpass filter (which is switched for each band). On receive, the signal goes through a highpass filter to get rid of broadcast band signals, and then travels through a bandpass filter which is also switched for each band. Notice that the BP filter is used on both transmit and receive. The signals go one way on transmit, and the opposite way on receive. This is done by some very clever diode switching. The final amplifier is permanently connected to the output. The transmit/receive relay is connected in the receive line, and is opened on transmit.

finger sensor

Down at the bottom of the block diagram are three blocks labeled DISPLAY, FREQUENCY SYNTHESIZED OSCILLATOR, and MICRO-PROCESSOR. They are NOT simple. The micro-processor is the device which makes everything happen in the HW-5400. It controls the display, the frequency synthesizer, and the finger sensor of the tuning knob. Yes, that's right . . . the finger sensor.

Now here's a bit of genius in design. The main



tuning knob has two finger-spinning holes. One of them has a metal insert. When your finger touches the metal insert, the two digits after the decimal on the display go out, and as you tune, the frequency increment is 1 kHz instead of 50 Hz. So when you want to make a big shift in frequency, use the metal insert spinner; for fine tuning, use the other one. On the unit that I built, I have the keypad accessory; for big jumps I use the keypad, and for 20 or 30 kHz, I use the spinner.

frequency control

The frequency synthesizer circuits control the VCOs, which combine to produce the required injection signal. The VCOs are controlled by a feedback loop referenced back to a crystal controlled oscillator. This loop has a bandpass filter, and the settling time for a VCO is inversely proportional to the bandwidth of the filter. The filter must be narrow enough to get rid of the reference frequency, and yet wide enough to allow quick response.

Heath solves the problem ingeniously. They set up the circuit so that the input to one of the divide-by-N chains is the difference between two VCOs. The step frequency of PLL 1 is 10.05 kHz; that of PLL 2, is 10 kHz. The difference between the two steps is only 50 Hz, yet the reference frequencies are at 10 kHz. This makes the filtering out of the reference frequency relatively easy. The output of VCO 1 is from 5.45-6.05 MHz (in 50 Hz steps). It is mixed with the BFO signal (8.83 MHz) to give 14.28-14.88 MHz. Then the signal is mixed with 10 or 20 MHz, and the output from VCO 3 or VCO 4, and the desired output frequency is produced. A lot of the switching is solid-state, so that the bandswitch is composed of only three wafers. One of those is used to control the solid-state switching circuitry.

The display on the HW-5400 is easy to read. At the left there is a series of symbols that indicate which mode the rig is in. There are several indicators — out of band, split frequency, transmit, and unlocked VCO. The VOX, anti-VOX, delay, and side tone controls are under a little flip-up panel to the right of the display.

When the Heath engineers laid out the printed circuit boards, they kept the transmit and receive parts pretty well separated, and generally speaking, the schematics follow the board layout in the same fashion. This makes it a little easier to locate the components on either the boards or the schematics.

operation

As I mentioned before, when I switched from one band to another, the HW-5400 remembered the frequency I last used on the original band. But it will also remember a second frequency. With eight bands (80, 40, 30, 20, 17, 15, 12, and 10), that makes a total of sixteen frequencies in memory, because the HW-5400 provides for split frequency operation on each band. Incidentally, when I operated split frequency, I found a slight delay in going from receive to transmit (but only on split frequency; simplex, the VCOs are already on frequency and there is no delay). This happened because the VCOs require a little time to settle.

I found it strange to be able to switch from one band to another without having to retune; in fact, I kept looking for the knobs. Fortunately, all of my antennas have fairly low SWR. At least they are low enough so that the HW-5400 will accept them without shutting down. Just to see what would happen, I put the rig on 75 meters and left the 20 meter beam antenna connected. The HW-5400 absolutely refused to put out any power. I didn't do any fancy tests to see at what SWR it would quit, but I figure that this is the mistake most likely to be made.

One disconcerting thing at first was "chirp" of the 50 Hz steps as I tuned in a CW or RTTY signal. But soon I no longer noticed the noise. For final tuning of signals, I started to use the RIT. It is smooth. Another feature I like is the IF shift. It seems to me that whenever I go off a net to talk to someone, the frequency I select invariably becomes the channel adjacent to the National Tuneup Frequency! With the IF shift, I can slide the IF sideways a little and get rid of the QRM. It's a nice feature.

While the high speed tuning feature is great, it did produce a surprise one evening. I spun the dial to a new frequency to meet someone for a QSO, and because Heath included a muting circuit to quiet the receiver while the VCO is settling, I ended up about 15 kHz past a local. He was transmitting when I tuned past him, but it took a few milliseconds before the speaker almost came off the desk from the buckshot. It took me a second or two to realize what had happened. I moved down another 15 or 20 kHz and learned something: the HW-5400 is almost as good as my tube rig for bearing up under close-by locals. I carried on the QSO about 30 kHz from the local, and although I knew he was there, it wasn't difficult or even uncomfortable to carry on the QSO.

Radio Amateurs are never satisfied. It doesn't much matter who designs what . . . we'd always like to see something different. For me, there are two things missing from the HW-5400: 160 meters and a way to connect up my SB-610 band scanner. After having used one for fifteen years or so, I feel almost lost without it connected.

One of the things I really like about this rig is the uncluttered front panel; you can tell what each control is for without searching through the instruction book. The knobs are far enough apart so that I don't feel crowded.

If you like to build, and have some experience

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at it, the HW-5400 is a good rig to tackle. If you want to dig inside to fix or tweak something, you'll be able to go ahead with confidence because you'll have been inside at least once before. You'll enjoy operating the HW-5400; it's a good rig.

For information on the HW-5400, contact Heath Company, Benton Harbor, Michigan 49022.

Fred Looker, VE3ZL

HW-5400: specifications

coverage:	80,40,30,20,17,15, 12,10 meters plus WWV at 10 MHz
readout:	7-digit display, vacuum fluorescent with special symbols
readout accuracy:	to nearest 50 Hz
frequency control:	synthesized
memory storage:	2 frequencies/band
stability:	less than 50 ppm drift from cold start
modes:	SSB normal and reverse CW — wide and narrow

receiver

sensitivity:	less than 0.35 μ V for 10 dB $\frac{S+N}{N}$
audio output:	2 watts in 4 ohms selectable (fast or slow)
intermod:	70 dB at 25 kHz

transmitter

RF output:	100 watts except on 10 meters (80 W)
load impedance:	90% rated power at 2:1 SWR high SWR protection
transmit-receive:	SSB: PTT or VOX CW — full break-in (simplex only)

general

dimensions:	11 1/2 x 14 x 5 inches (29.2 x 35.6 x 12.7 cm)
weight:	24 pounds (10.9 kg)
power:	13.8 volts DC at 20 amperes (HWA-5400-1 power supply: 120/240 V, 60 Hz)

**book review:
radio frequency design**

Wes Hayward, W7ZOI, an engineer from Tektronix, is well known to the Amateur and professional community. His numerous efforts to educate designers in both detailed schematics and system approach have led him to write *Introduction to Radio Frequency Design*, published by Prentice Hall in January, 1982.

Having seen many books on this subject, I was impressed with the way this one was composed and the way issues are addressed. The book consists of eight chapters covering low frequency transistor models, filter basics, coupled resonator filters, transmission lines, two-port networks, amplifiers and mixers, and oscillators and frequency synthesizers. A final chapter is titled "The Receiver: An RF System."

The material presented is complete; references are well chosen. The book reads easily and makes one feel sufficiently impressed to keep on reading. I was tempted to use my computer resources to verify some of the more complex numerical examples, and I was not surprised to find that the author's calculations were correct. I especially enjoyed a few specific items (his non-linear model of the transistor and his analysis of noise, for example) and recommend these sections — and the book as a whole — to all readers. *Introduction to Radio Frequency Design* is thoroughly enjoyable and well worth the \$29.95 investment.


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Sherwood filters

Receiver selectivity and the ability to reject unwanted signals are subjects of tremendous interest to Radio Amateurs. Long known for their excellent aftermarket filters and extensive modification of the R4C receiver, Sherwood Engineering now has a line of front end antenna filters. My interest came as a result of a phone call from George Heidelberg at Sherwood alerting me to a new filter they had designed recently for the 160 meter DX window, 1.825-1.830 MHz.

If you haven't been on 160 meters recently, you would be surprised by the amount of activity that will be found. On almost any given evening there are plenty of CW and SSB QSO's in progress; rapid growth and the elimination of power restrictions has resulted in some crowding on the band. For instance, SSB operators who sit just above 1.830 MHz with an LSB signal put a fair part of their signal down into the DX window.

During both phone and CW contests, when 160 can sound as bad as 20 meters, the DX window is often bracketed by strong signals. For the average transceiver or receiver, this can result in very difficult copy in the DX window area. Installation of the Sherwood filter will pass only these signals that fall between 1.825 and 1.830 MHz thus facilitating easier reception without front-end overload, intermodulation products, etc. Another suggested application is at large multi-op, multi-transmitter contest stations; a spotter can scan the DX window, minimizing interference, while the operator is working stations on another part of the band. When another multiplier is heard, the operator can QSY to the appropriate frequency and attempt to work the station.

The filters can also be used to reduce interference from high powered shortwave broadcast stations and in high ham density, urban environments. Currently Sherwood Engineering has available filters that cover 25 kHz segments for the 40 and 20 meter CW bands, 12.5 kHz segments of 80/75 meters, and any 5 kHz segment on 160 meters. Other frequencies are available on request.

These low-loss front-end antenna filters are based upon a high performance, 6 pole, crystal 50 ohm design. They have a shape factor of 2.25:1, 6/60 dB and are designed to be used with any receiver and can be adapted to most transceivers.

the acid test

After receiving the filter from Sherwood, I reconfigured my receiver input so that I could switch in and out the filter to do A-B comparisons. Because I live in a low-density area, having other stations nearby is not a problem. However, as more and more new hams have joined the fun on 160 meters, the band has become significantly more crowded. During several 1983 contests, I found that while listening with the filter in, I was able to effectively reduce interference coming from other stations outside

the DX window. At the filter's edges, nearby signals were not completely eliminated due to operating at the 3 dB points. My overall impression of these front-end filters is that they would be highly desirable at a multi-op, multi-transmitter operation and in any high ham density area. For 160 meters, they provide an extra measure of selectivity to help ferret out those weak and hard-to-copy DX signals. They can also be used ahead of a Beverage antenna pre-amplifier.

Filters for several frequency segments are available for each band. They include the FE-14000/6, FE-14200/6, FE-7000/6, FE-3500/6, and FE-1825/6, as well as others that may be specially ordered. Most are priced at \$80; 160 meter filters are priced at \$145. Add \$3 for domestic shipping, \$6 for shipping overseas.

For more information on Sherwood's front-end filter or any of the rest of their product line, contact Sherwood Engineering, 1268 S. Ogden Street, Denver, Colorado 80210.

Circle #302 on Reader Service Card.

— N1ACH



remote control antenna tuners

Here's a brand new product of interest to hams who've been disappointed with their antenna's performance. The VT-3/VT-4 is a remotely tuned series-fed capacitor that is connected directly to the antenna. Adjusting the VT-3/VT-4 will tune the antenna to a minimum value of SWR. For example, mobile antennas that would before only tune 20 kHz without re-adjustment, will now be able to tune the whole band by a flick of the switch. Other suggested uses are with a trapped vertical, half wave doublet, long wire, or sloper.

The VT-3/VT-4 is housed in an aluminum universal base mount with an anodized aluminum cover for weather protection. With the addition of the LC-4 inductor kit, it can be converted to a voltage-fed matching device. This is a single band system that can cover 1.8-30 MHz and will provide full size band coverage with appropriate coil tap adjustment.

The VT-3 is designed for mobile installation and operates directly from the vehicle's 12-volt power supply. The VT-4 is a 117 VAC - 12 VDC supply and switch. The VT-3/VT-4 has a switch to activate the capacitor and limit lights show when either maximum or minimum capacitance values are reached. Both units require eight conductor control cables. Current consumption is low so cable runs can be long.

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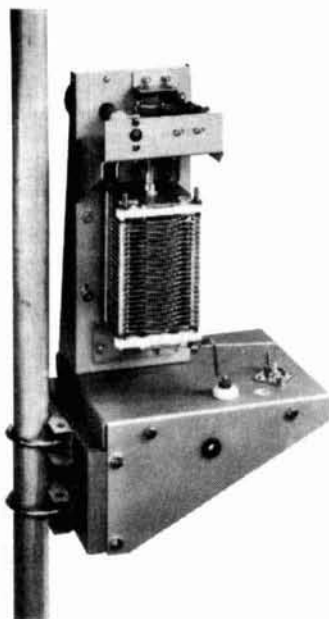
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NEW products

Price for the VT-3/VT-4 is \$159.95. Eight conductor cable is available for 24 cents per foot and the LC-4 inductor kit is priced at \$39.00. All Vector Radio products are sold with a money-back guarantee.



For more information on the Vector Radio VT-3/VT-4 remote antenna tuner, contact Vector Radio Co., P.O. Box 1166, Cardiff, California 92007.

Circle #303 on Reader Service Card.

radio modem

Macrotronics, Inc. has introduced the RM1000 radio modem, a hardware and software system that converts a personal computer into a state-of-the-art communications terminal supporting Morse, Baudot and ASCII codes. The radio modem is intended for use by Amateur Radio operators and SWL's for copying news and wire services. It features commercial quality demodulators, dual bargraph tuning, and extensive software capabilities.

The RM1000 uses multistage active filter demodulators with dual LED bargraph tuning indicators for reception of both Morse code and radioteletype (RTTY) signals. It offers three RTTY shifts which may be selected from the computer keyboard, and a crystal-controlled

AFSK tone generator provides stable RTTY keying. Relays are used for Morse code and push-to-talk transmitter keying. A hardware clock continuously displays time and may be inserted into text in a completely user-programmable format.

Many features are included in the software to accommodate a wide variety of operating situations including net operations, MARS, RTTY, art ("PIX"), contesting and SWLing. A 70-page user manual is included.



The RM1000 radio modem system is currently available for ATARI™, APPLE™, IBM™ and Radio Shack TRS80™ microcomputers. For complete information, contact Macrotronics, Inc., 1125 N. Golden State Blvd., Turlock, California 95380.

Circle #304 on Reader Service Card.

new solid-state tube for Drake R-4

Sartori Associates has just announced the availability of a new solid-state tube, the SBA6. Designed to replace the 6BA6 in Drake R-4(A-B-C) receivers, the SBA6 will also replace the RF and IF 6BZ6 vacuum tubes in the R-4A/B and early model R-4C, as well as the 6BA6 and 12BA6 vacuum tubes used in the IF amplifiers of the R-4 series. (For the third mixer in the early model R-4C, we recommend replacing the 6BA6 with a 6HS6/SHS6 — (Drake made this improvement in the mid-model R-4C). The new SBA6 will also serve as a plug-in replacement for your T-4X 12BA6 ALC'd IF amplifier.

Sartori solid-state tubes provide no-warm-up, high performance, trouble-free operation with R-4(A-B-C) receivers. Sartori also manufactures SEJ7, SHS6, and SBE6 mixers for the R-4 series and the SEJ7, SHS6, SAU6, SAX7-1, SAX7-2, SEV7/SFQ7/SAQ8 for the T-4X series. All are priced at \$23.00 postpaid.

For more information about the SBA6, contact Sartori Engineering, P.O. Box 2083, Richardson, Texas 75080.

Circle #305 on Reader Service Card.

IC-04A and IC-04AT

The IC-04A and IC-04AT, two new 440 MHz HTs from ICOM, feature frequency entry, con-

trol functions and 32 PL tones controlled by the 16-button pad on the face of the radio. Also in-



cluded are priority, scanning (both of memories and programmable band scan), and DTMF (04AT only). For scanning, 5 kHz increments are front-panel selectable. Ten memories with internal lithium battery backup give the ultimate in flexibility for channelizing operation for easy access to most-used channels. A custom LCD readout with S-meter is unique to the ham industry.

The IC-04A and IC-04AT have the same styling, control features and functions of the IC-02A(T), and utilize the existing accessory line available for the IC-2A and IC-2AT, plus new accessories such as long-life and high-power battery packs.

For details, contact ICOM, 2112 116th Avenue, N.E., Bellevue, Washington 98004.

Circle #308 on Reader Service Card.

full-function DTMF decoder

A full-function, dual-tone multifrequency decoder module, model 2009, is a state-of-the-art CMOS design which decodes all 16 DTMF codes. Available from Proham Electronics, model 2009 has several advanced features such as a crystal-controlled timebase for long-term accuracy, on-board voltage regulation, counter detection with period averaging to minimize falsing, and latched 4-bit digital outputs with a choice of binary or row/column format.

Two LSI CMOS chips provide high performance operation and minimize the parts count. All bandpass and band reject filtering is achieved by using one switched-capacitive filter integrated circuit. The time base oscillator and dividers are also within this chip. Likewise, the actual decoding is performed by the second CMOS device. This simplifies operation, since there is only one adjustment required, and it sets the operating level. There are no frequency adjustments because all timing is referenced to the crystal controlled time base. Ancillary benefits resulting from the application of these LSI CMOS devices is compact size, the printed circuit board is only 3.6 x 2.0 inches, and low power requirement, typically 40 mA at 12.5 VDC. The kit is easy to

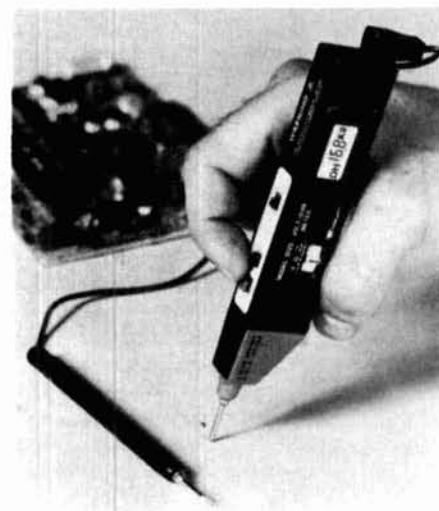
build using the comprehensive instruction manual supplied, and easy to use. When driven with an audio signal between 50 mV and 1.0 V, the model 2009 produces a 4-bit digital output code corresponding to the DTMF digit detected and valid code pulse. This output can be used to drive a parallel port of a microcomputer or additional digital logic circuitry as required. The price of a bare board with manual is \$9.95; board, manual, filter and decoder chip, \$44.95; complete kit, \$99.95; and manual only, \$5.00. All prices include postage in U.S.A. (Ohio residents add 5% sales tax.)

For further information, contact Proham Electronics, Incorporated, 34620 Lakeland Blvd., Eastlake, Ohio 44094.

Circle #306 on Reader Service Card.

shirt-pocket volt-ohmmeter

The new Model 3525 DIGI-PROBE™ volt-ohmmeter, just introduced by Triplett, is said to be one of the smallest trouble-shooting, battery-operated, digital instruments presently manufactured. Its shirt-pocket size, internal overload protection, accuracy and auto-ranging features make it appropriate for a myriad of



lab, circuit design or in-field measurements on industrial, commercial, or consumer electronic/electrical equipment.

Only 6-3/8" long x 1-1/8" wide x 3/4" deep (162 x 28 x 20 mm), the DIGI-PROBE™ utilizes a large 5mm easy-reading 3.5 digit LCD display with a convenient "Data Hold" feature to facilitate measurements in low ambient light or in confined areas. It also enables the user to "hold" the reading for later review. An instant-continuity test permits rapid testing of diodes, shorts and circuit continuity. Volts, ohms, and continuity are easily selected with a simple function switch. AC and DC volts are selected by push-button with AC shown on the LCD display.

Auto-ranging on volts and ohms functions

eliminates the need for range selection, providing true "Touch and Test" capability. The Model 3525 has thirteen ranges. Range selection in all functions is fully automatic. The ranges are: 0-500 VDC in four ranges; 0-500 VAC in four ranges; 0-2.0 Megohms (2.9 Megohms in overrange) in four ranges. Auto-ranging response time is 5 seconds maximum and accuracy is ±0.75 percent of RDG + two digits on most ranges. Blinking-digit overrange indication and low battery visual indication are provided. Internal overload protection is to 750 VAC/DC in voltage ranges and 250 VAC/DC in ohms and continuity ranges.

The DIGI-PROBE™ case is molded of high impact black thermoplastic with textured surface. The unit weighs only 2-1/2 ounces (0.07 kg).

Priced at \$65.00, the DIGI-PROBE™ is furnished with two 1-1/2 volt button-type batteries, shirt-pocket carrying case, attached 28-inch test lead, comprehensive instruction manual and one-year warranty.

For information or a free demonstration of the DIGI-PROBE™ contact Triplett Corporation, One Triplett Drive, Bluffton, Ohio 45817.

Circle #309 on Reader Service Card.

noise figure measurements application note

HP's new application note AN/57-1, "Principles of RF and Microwave Noise Figure Measurement," is now available for all those working on device, component, sub-system and system noise figure. It replaces the long-popular AN/57, "Noise Figure Primer."

The 40-page note serves as a comprehensive tutorial on noise figure, with detailed material on thermal and shot noise, concepts of noise figure, effective noise temperature, Y-factor, etc.

Plenty of useful information is provided on subtle measurement considerations including single sideband vs. double sideband, effects of local-oscillator noise, second-stage effects and corrections, hot/cold techniques, frequency conversation and image considerations.

An extensive glossary includes common symbols and detailed technical explanations of most terms. Also included is a bibliography of 34 other noise-figure-related references.

For a free copy of AN/57-1, contact Hewlett-Packard Company, 1820 Embarcadero Road, Palo Alto, California 94303.

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COUNSELORS: Connecticut brother-sister camp. Completely equipped with ham radio station. Program includes electronics, kit building, code and communications. June 25-August 22. Send resume: Lloyd Albin (N2DMQ) Ken-Mont and Ken-Wood Camps, 2 Spencer Place, Scarsdale, NY 10583.

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NEED TO CONTACT James Navarchi concerning Yaesu gear. C. T. Huth, 146 Schonhardt St., Tiffin, OH 44883

FASTRAK 2005 Active filter features 8-pole response and dc switching, make a CW filter with 80 hertz bandwidth; 3.6 x 3.0 inch pcb and manual \$9.95. Proham Electronics, Inc., 34620 Lakeland Blvd., Eastlake, OH 44094.

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WANTED: Manual for Lavoie oscilloscope model LA-265A. Original or photocopy. State cost. WE4O, 4257 Via Alta Drive, Mobile, Alabama 36609.

SALE: Vanguard Labs signal generator 100 to 180 MHz, Johnson CB test set. Checks power, modulation and receiver. Jeff V. Nielson, 610 South 200E, Monroe, Utah 84754. (801) 527-4242.

CABLE CONVERTERS, decoders. Catalog \$1 refundable. APS, POB 263 HR, Newport, RI 02840

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TENNESTEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed. Send stamp for details. WBURR, 1025 Wildwood Road, Quincy, MI 49082.

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WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

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VERY in-ter-est-ing! Next 4 issues \$2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

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WANTED: ARRL VHF Manual, early edition. State condition and year of publication. Stanley Jaffin, WB3BGU, 800 Stonington Road, Silver Spring, Maryland 20902. Telephone 301 593-9599, evenings.

Coming Events ACTIVITIES "Places to go..."

OHIO: The Cuyahoga Falls ARC's 30th annual Electronic Equipment Auction and Hamfest, Sunday, February 26, North High School, Akron. 8 AM to 4 PM. Tickets \$2.50 advance, \$3.00 at door. Sellers may bring own tables or some available for \$2.00, advance table reservations advised. Talkin on 87/27. For information: CFARC, P.O. Box 6, Cuyahoga Falls, OH 44222 or call KBJSL (216) 923-3830.

INDIANA: The LaPorte Amateur Radio Club's Winter Hamfest, Sunday, February 26, Civic Auditorium in LaPorte. Starts 7 AM Chicago time. Admission \$2.50 per person. 8 ft. long tables available for \$2.00 each by reservation. Good food, coffee, etc. Talk in on 52 simplex. SASE for tables, tickets, or information to LPARC, P.O. Box 30, LaPorte, IN 46350.

MICHIGAN: The 14th annual Livonia Amateur Radio Club's Swap 'n Shop, Sunday, March 4, 8 AM to 4 PM, Churchill High School in Livonia. Plenty of tables, refreshments and free parking. Talk in on 144.75/35 and 52 simplex. For further information send large SASE to Neil Coffin, WA8GWL, Livonia ARC, P.O. Box 2111, Livonia, MI 48151.

OHIO: Cincinnati ARRL '84 State Convention and Flea Market, February 25 and 26. Registration \$5. Flea market \$4 per space both days. Forums, meetings, vendors, Wouff Hong, banquet. Hospitality suite Friday and Saturday nights. Write: Cincinnati ARRL '84, POB11300, Cincinnati, OH 45211 or call (513) 825-8234.

KENTUCKY: Glasgow Swapfest, Saturday, February 25, 8 AM CST til... Glasgow Flea Market Building 2 miles south of Glas-

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NEW JERSEY: Springfest '84 sponsored by the Shore Points ARC, Saturday, March 10 from 9 AM to 4 PM, Atlantic County 4-H Center, Egg Harbor City, approx. 15 miles west of Atlantic City. Large heated, indoor selling space. Covered tailgating (weather permitting). Sellers \$5 per space with own table. Buyers \$2.50 advance, \$3.00 day of hamfest. For information, SPARC, P.O. Box 142, Absecon, NJ 08201.

PENNSYLVANIA: The 1984 Lancaster Hamfest, Sunday, February 19, Guernsey Sales Pavilion, U.S. Rts. 30 and 896, Lancaster. 0800 to 1600, dealer setup 0600. Commercial tables (main hall) \$15.00. Non-commercial (rear annex) \$6.00. General admission \$3.00. Tailgating free with general admission (weather permitting). Talk in on 146.61 and 147.015. Send reservations to Hamfest Committee, P.O. Box 6082, Lancaster, PA 17603. Please make checks payable to SERCOM, Inc.

MICHIGAN: The Cherryland Amateur Radio Club's 11th annual Swap N Shop, February 11, Immaculate Conception Elementary School gym, 218 Vine Street, Traverse City, 8 AM to 2:30 PM. Tables \$3.00 each with setup at 6:30 AM. Admission \$2.50. Talk in on 146.25/85. For details SASE to Jerry Cermak, K8YVU, 3905 Slusher Rd., Traverse City, MI 49684.

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OPERATING EVENTS

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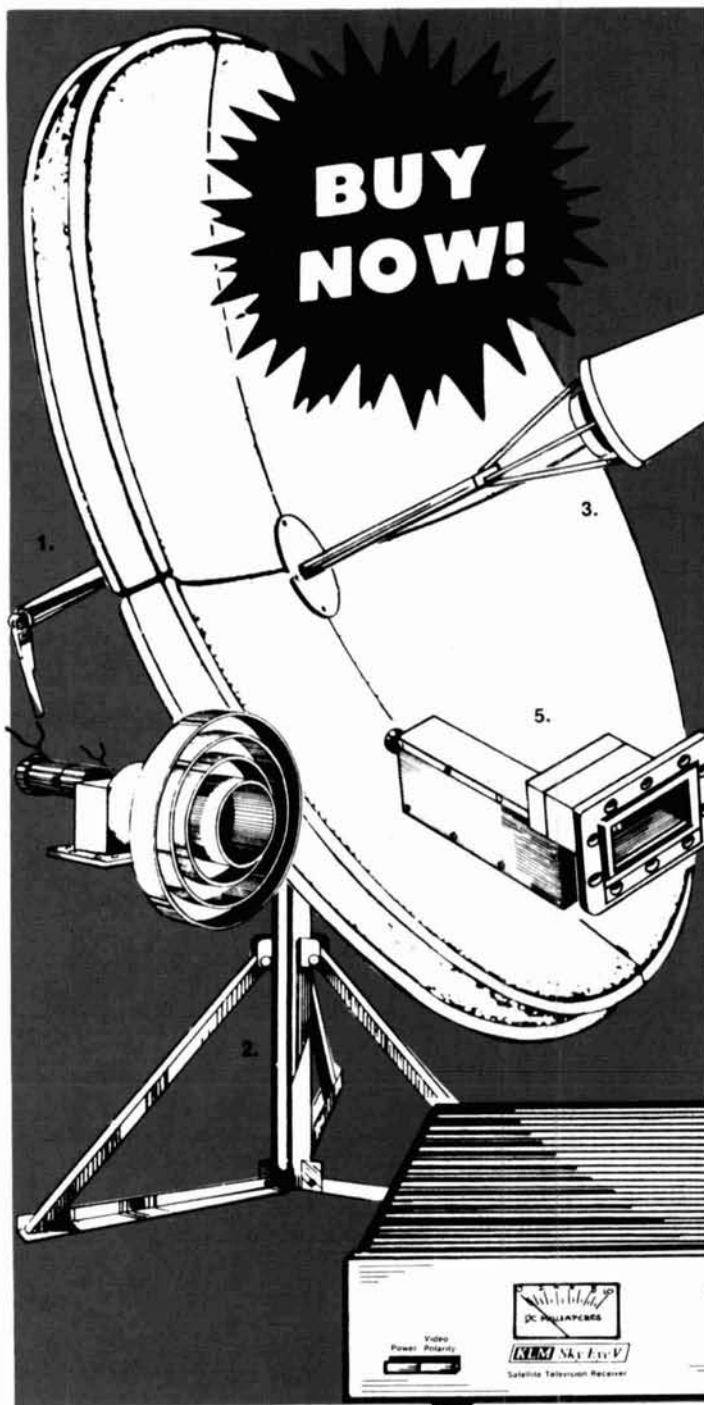
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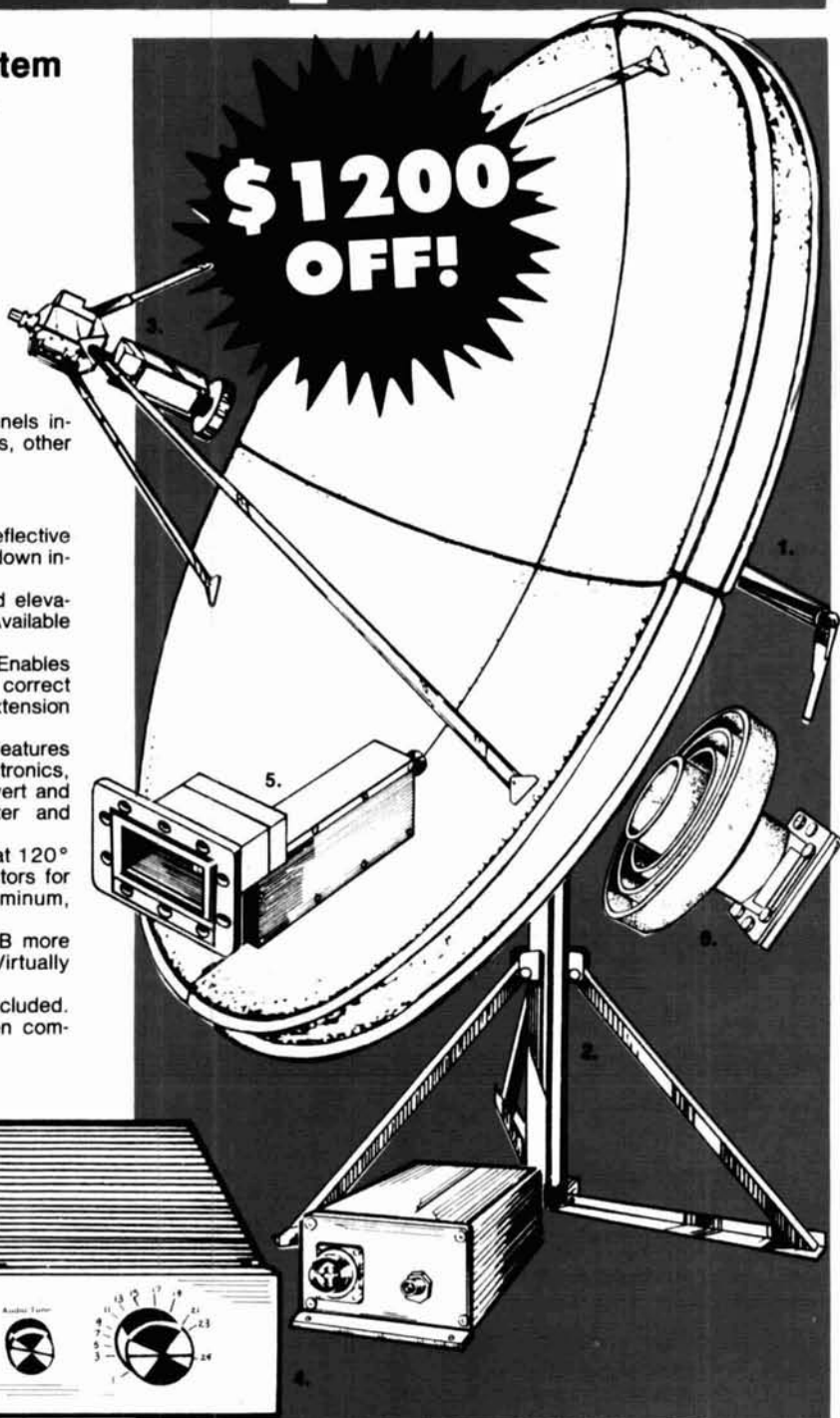
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Receives up to 60 satellite TV channels including movies, sporting events, news, other TV stations and much more!

What the system includes:

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2. Polar mount. Features azimuth and elevation adjustments for precise tracking. Available with earth or slab mounts.
3. LNA mount complete with rotor. Enables LNA to be remotely positioned for correct polarity. Control console included. Extension tubes extra.
4. KLM Sky Eye V satellite receiver. Features the latest in single conversion electronics, large signal strength meter, video invert and variable audio tuning. Downconverter and built-in RF modulator included.
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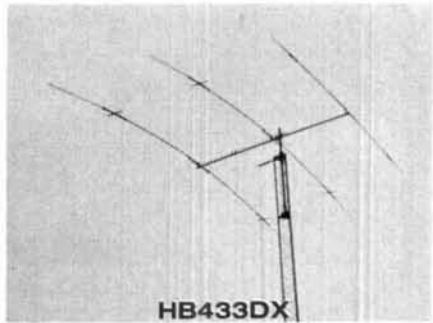
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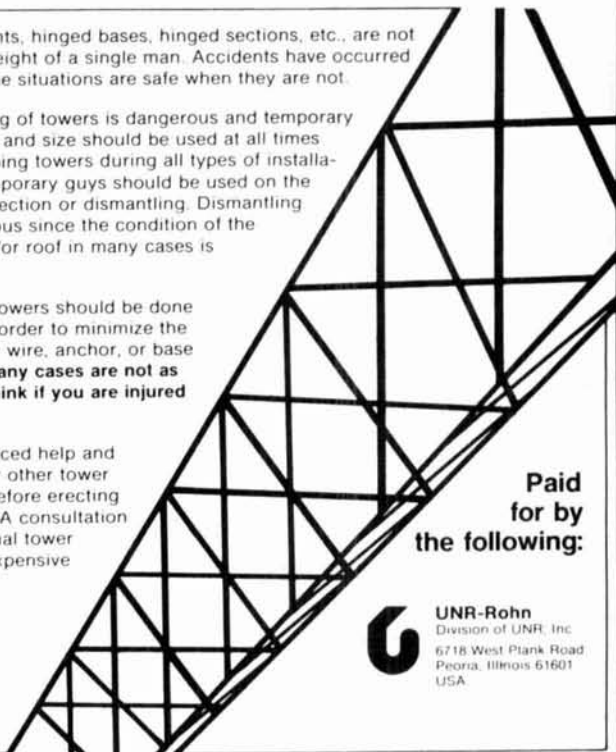
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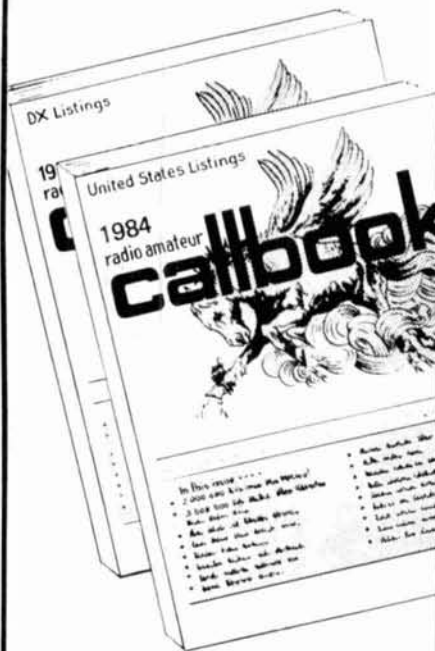
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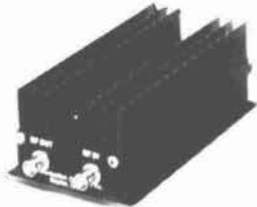
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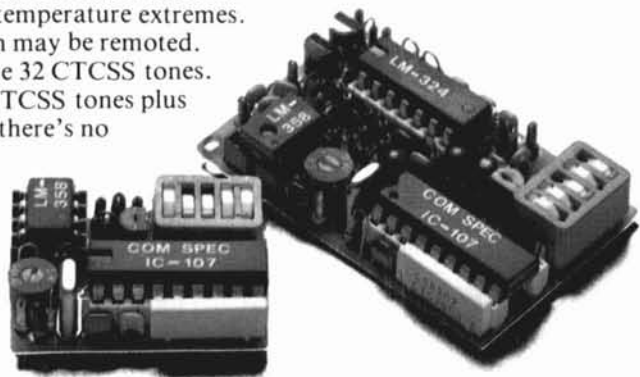


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R-2000 FEATURES:

- **Covers 150 kHz-30 MHz in 30 bands.** Uses innovative UP-conversion digitally controlled PLL circuit. UP/DOWN band switches (1-MHz step). VFO's continuously tuneable across the band and from band to band.
- **Optional 118-174 MHz coverage.** Through use of innovative microprocessor technology, frequency, band, and mode data of stations in the 118-174 MHz range may be tuned, displayed (full frequency, i.e., 146.000.0), stored in memory, recalled, and scanned, using the R-2000 front panel controls and frequency display, allowing maximum convenience and ease of operation.
The optional VC-10 VHF converter unit may be easily installed on the rear panel of the R-2000.
- **All mode: USB, LSB, CW, AM, FM.** Provides expanded flexibility in receiving various signal types. Front panel mode selector keys, with LED indicators.
- **Digital VFO's for best stability.** 50-Hz step, switchable to 500-Hz or 5-kHz. F. LOCK switch provided.
- **Ten memories store frequency, band, and mode data.** Complete information on frequency, band, and mode is stored in memory, assuring maximum ease of operation. Each memory may be tuned as a VFO. Original memory frequency may be recalled. AUTO. M switch for automatic storage of current operating data, or, when off, selective storage of data using M. IN switch.
- **Lithium battery memory back-up. (Est. 5 yr. life.)**
- **Programmable memory scan.** Scans all memories, or may be programmed to scan specific memories. HOLD switch interrupts scanning. Frequency, band, and mode are automatically selected in accordance with the memory channel being scanned. The scanning time is approximately 2 seconds per channel.
- **Programmable band scan.** Scans automatically within the programmed bandwidth. Memory channels 9 and 0 establish upper and lower scan limits. HOLD switch interrupts scanning. Frequency may be adjusted, using the tuning control, during scan HOLD.
- **Fluorescent tube digital display (100-Hz resolution).** Built-in 7 digit fluorescent tube digital display indicates frequency or time, plus memory channel number. DIM switch provided. The display may be switched to indicate CLOCK-2, FREQUENCY, CLOCK-1, and timer ON or OFF by the front panel FUNCTION switch.
- **Dual 24-hour quartz clocks, with timer.**
- **Three built-in IF filters with NARROW/WIDE selector switch. (CW filter opt.)** 6-kHz wide or 2.7-kHz narrow on AM. 2.7-kHz automatic on SSB. 2.7-kHz wide on CW, or, with optional YG-455C filter installed, 500-Hz narrow. 15-kHz automatic on FM.
- **Squelch circuit, all mode, built-in, with BUSY indicator.**
- **Noise blanker built-in.**
- **Large front mounted speaker.**
- **Tone control.**
- **RF step attenuator. (0-10-20-30 dB.)** Four step attenuator, plus antenna fuse.
- **AGC switch. (Slow-Fast.)**
- **"S" meter, with SINPO "S" scale.**
- **100/120/220/240 VAC, or 13.8 VDC operation** (with opt. DCK-1 cable kit).

Other features.

- RECORD output jack.
- Audible "beeper" (through speaker).
- Carrying handle.
- Headphone jack.
- External speaker jack.

Optional accessories:

- VC-10 118-174 MHz converter.
- HS-4, HS-5, HS-6, HS-7 headphones.
- DCK-1 DC cable kit.
- YG-455C 500-Hz CW filter.
- HC-10 World digital quartz clock.
- AL-2 Surge Shunt

More information on the R-2000 is available from all authorized dealers of Trio-Kenwood Communications
1111 West Walnut Street
Compton, California 90220.

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